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[54] METHOD AND APPARATUS GENERATING HIGH RESOLUTION DATA AND ECHO IDENTIFICATION

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[51] Int. Cl.[5] G03B 42/06

[52] U.S. Cl. 367/7; 367/11; 128/660.01

[58] Field of Search 367/7, 11, 87, 135; 364/421; 342/192; 128/660.01; 73/602, 607, 620, 627

[56] References Cited

U.S. PATENT DOCUMENTS

4,862,180 8/1989 Martin 342/192

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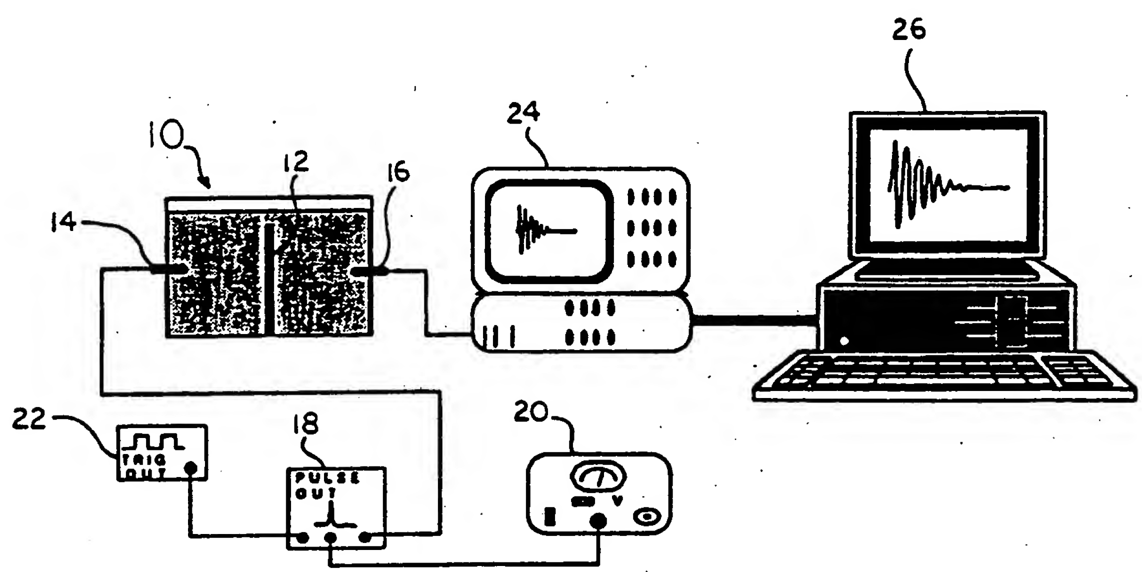
Attorney, Agent, or Firm—Oldham & Oldham Company

[57] ABSTRACT

The present invention is generally directed to an apparatus and method of echo identification wherein the sources of echoes generated from interface boundaries in a medium may be more effectively identified in the processing of obtained waveform data to generate higher resolution in the obtained data. The method of the present invention is particularly useful for performing ultrasound reflectometry and medical sonography by obtaining data representative of a medium being studied and suppressing signals from overlapping echo waveforms generated from obstacles adjacent an object under test. More particularly, the method of the invention enables the sources of such echo waveforms to be more effectively identified to yield an indication of the contribution of overlapping echoes to generate higher resolution data signals. The processing includes generating a power spectral estimate of the data which is utilized to obtain a phase estimate thereof. The Maximum Entropy Estimation Method is used to obtain a reliable phase estimate which can then be processed to yield an indication of the delay time to an interface boundary in the medium. The sources of the echo waveforms can then be located and used to increase the resolution of the obtained data.

21 Claims, 10 Drawing Sheets



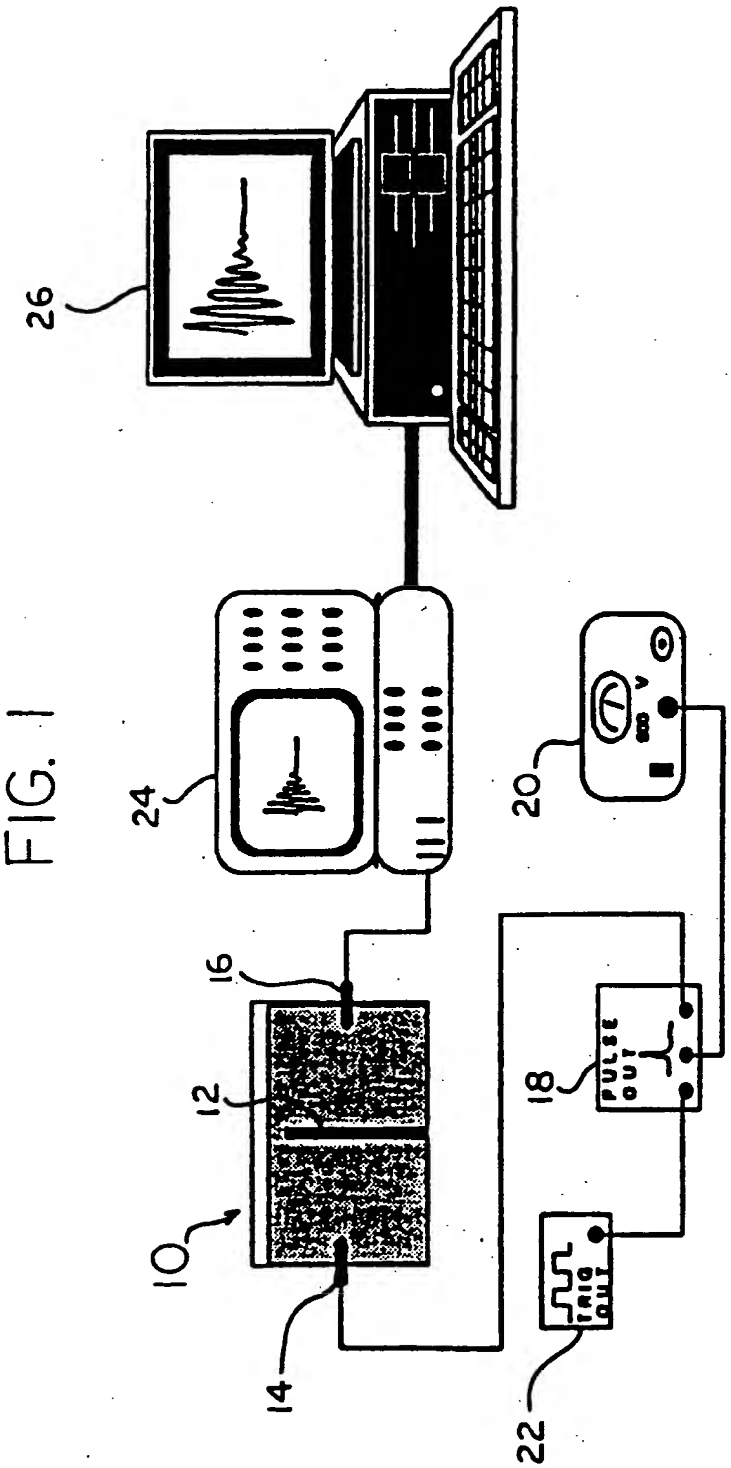


FIG. 1

FIG. 2

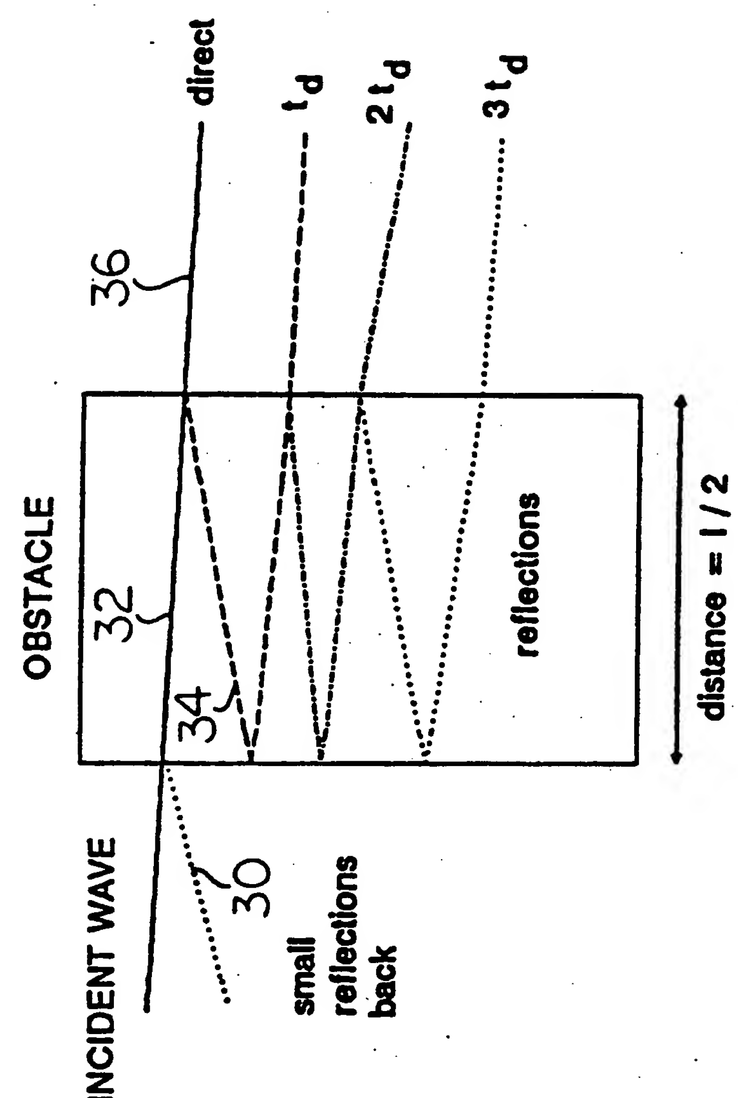

OBSTACLE
INCIDENT WAVE
direct
t d
2 t d
3 t d
36
32
34
30
small reflections back
reflections
distance = l / 2

FIG. 3

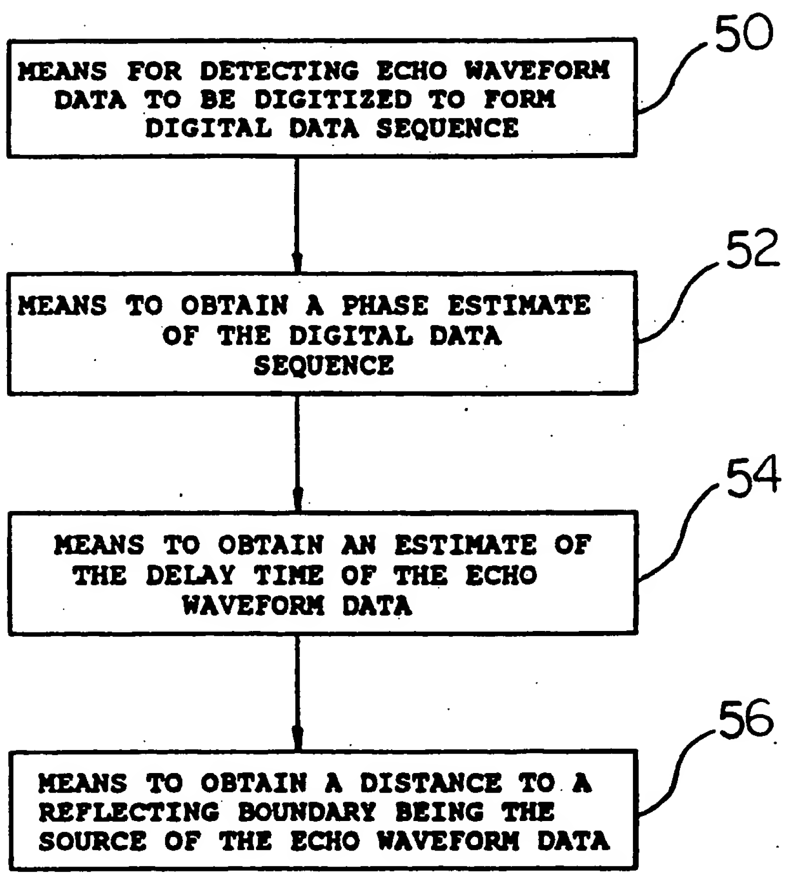
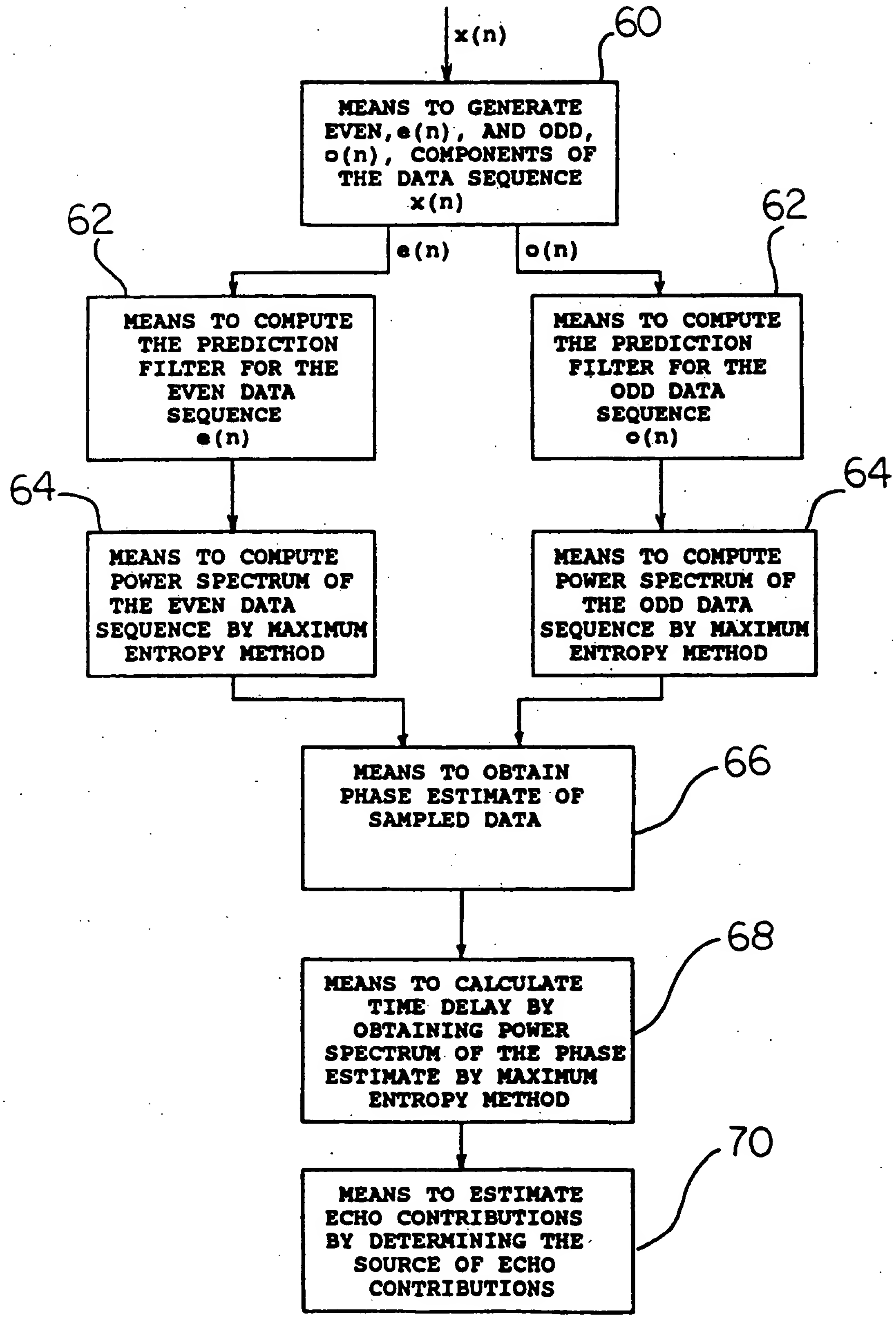

MEANS FOR DETECTING ECHO WAVEFORM DATA TO BE DIGITIZED TO FORM DIGITAL DATA SEQUENCE
50
MEANS TO OBTAIN A PHASE ESTIMATE OF THE DIGITAL DATA SEQUENCE
52
MEANS TO OBTAIN AN ESTIMATE OF THE DELAY TIME OF THE ECHO WAVEFORM DATA
54
MEANS TO OBTAIN A DISTANCE TO A REFLECTING BOUNDARY BEING THE SOURCE OF THE ECHO WAVEFORM DATA
56

FIG.4


x(n)
60
MEANS TO GENERATE EVEN, e(n), AND ODD, o(n), COMPONENTS OF THE DATA SEQUENCE x(n)
e(n)
o(n)
62
MEANS TO COMPUTE THE PREDICTION FILTER FOR THE EVEN DATA SEQUENCE e(n)
62
MEANS TO COMPUTE THE PREDICTION FILTER FOR THE ODD DATA SEQUENCE o(n)
64
MEANS TO COMPUTE POWER SPECTRUM OF THE EVEN DATA SEQUENCE BY MAXIMUM ENTROPY METHOD
64
MEANS TO COMPUTE POWER SPECTRUM OF THE ODD DATA SEQUENCE BY MAXIMUM ENTROPY METHOD
66
MEANS TO OBTAIN PHASE ESTIMATE OF SAMPLED DATA
68
MEANS TO CALCULATE TIME DELAY BY OBTAINING POWER SPECTRUM OF THE PHASE ESTIMATE BY MAXIMUM ENTROPY METHOD
70
MEANS TO ESTIMATE ECHO CONTRIBUTIONS BY DETERMINING THE SOURCE OF ECHO CONTRIBUTIONS

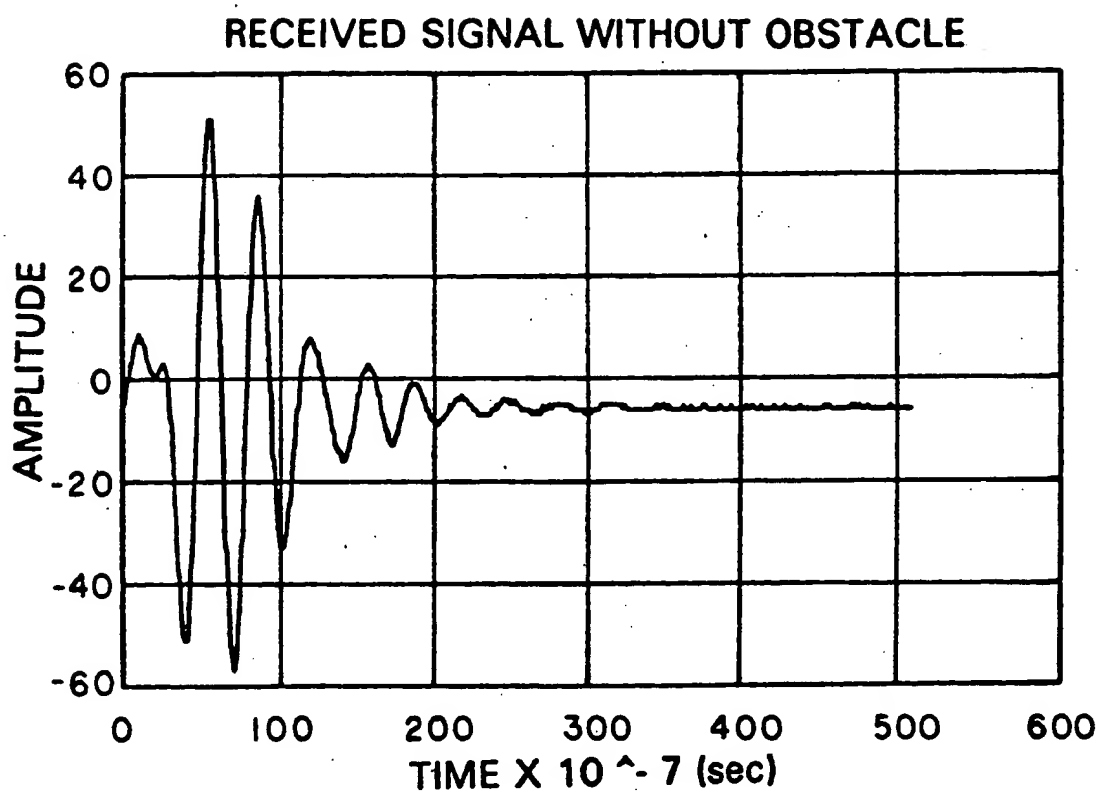

RECEIVED SIGNAL WITHOUT OBSTACLE
60
40
20
0
-20
-40
-60
AMPLITUDE
0
100
200
300
400
500
600
TIME X 10 ^- 7 (sec)

FIG.-5

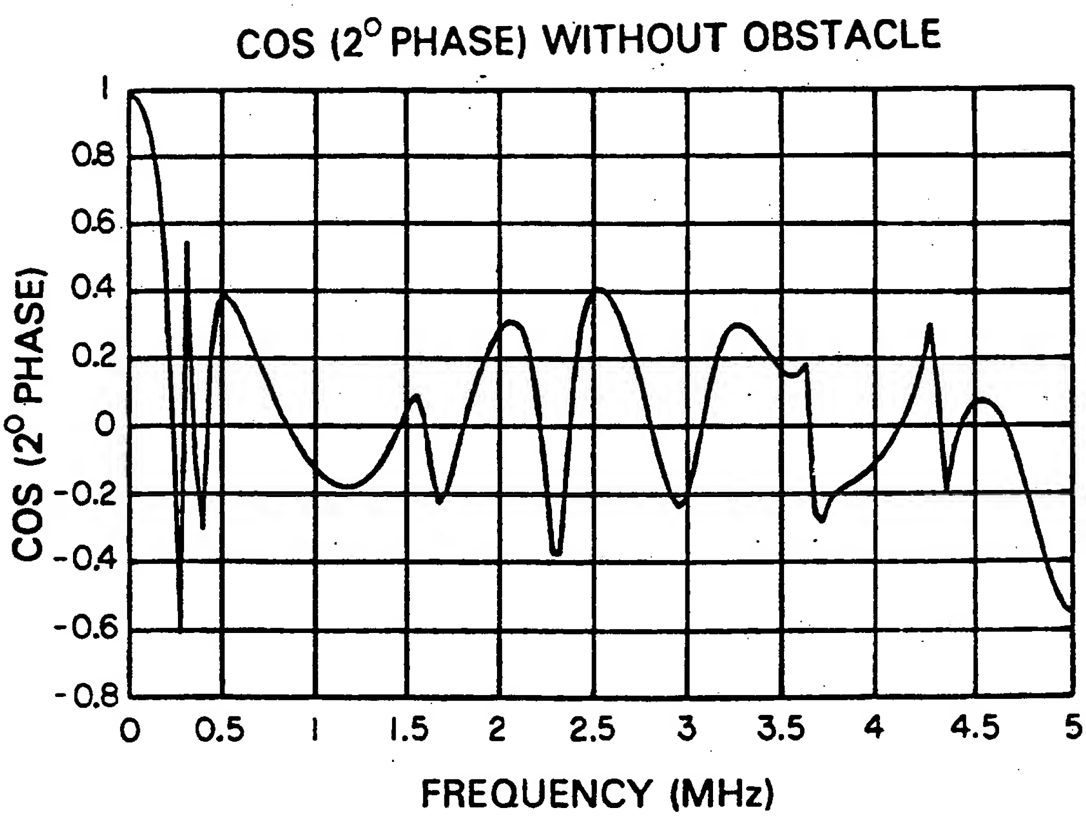

COS (2° PHASE) WITHOUT OBSTACLE
1
0.8
0.6
0.4
0.2
0
-0.2
-0.4
-0.6
-0.8
COS (2° PHASE)
0
0.5
1
1.5
2
2.5
3
3.5
4
4.5
5
FREQUENCY (MHz)

FIG.-6

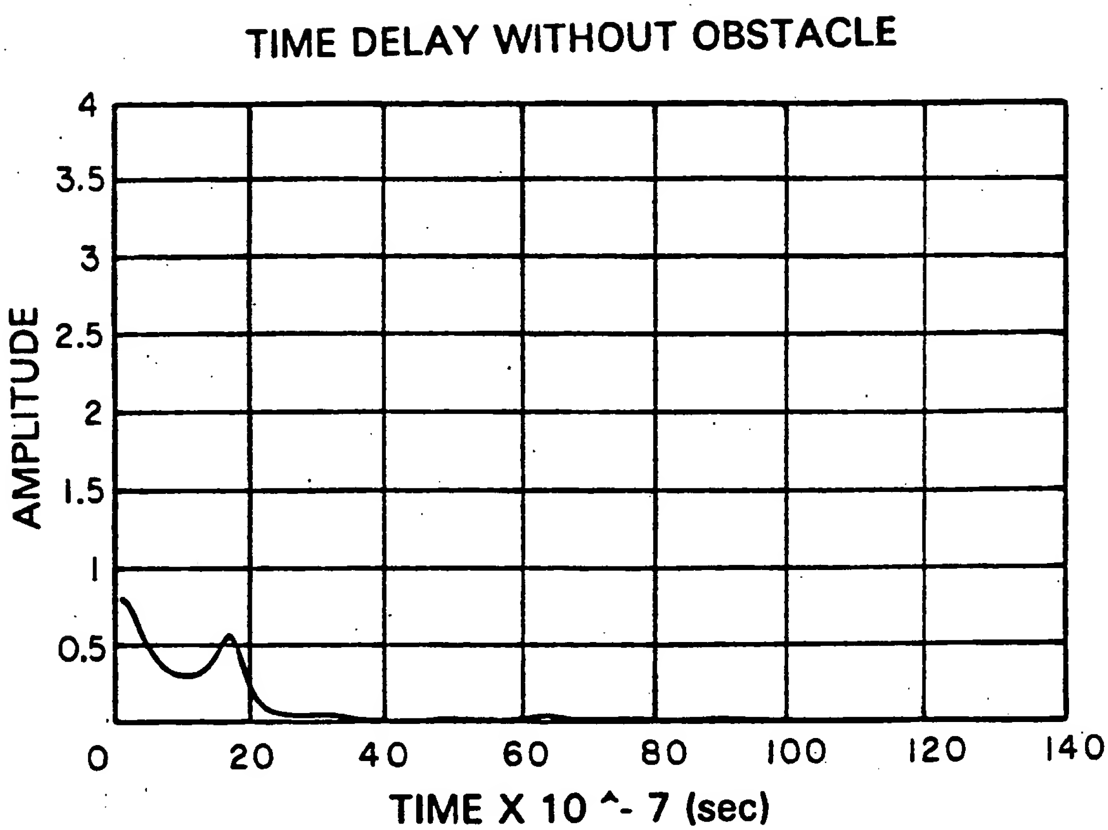

TIME DELAY WITHOUT OBSTACLE
AMPLITUDE
4
3.5
3
2.5
2
1.5
1
0.5
0
20
40
60
80
100
120
140
TIME X 10 ^- 7 (sec)

FIG.-7

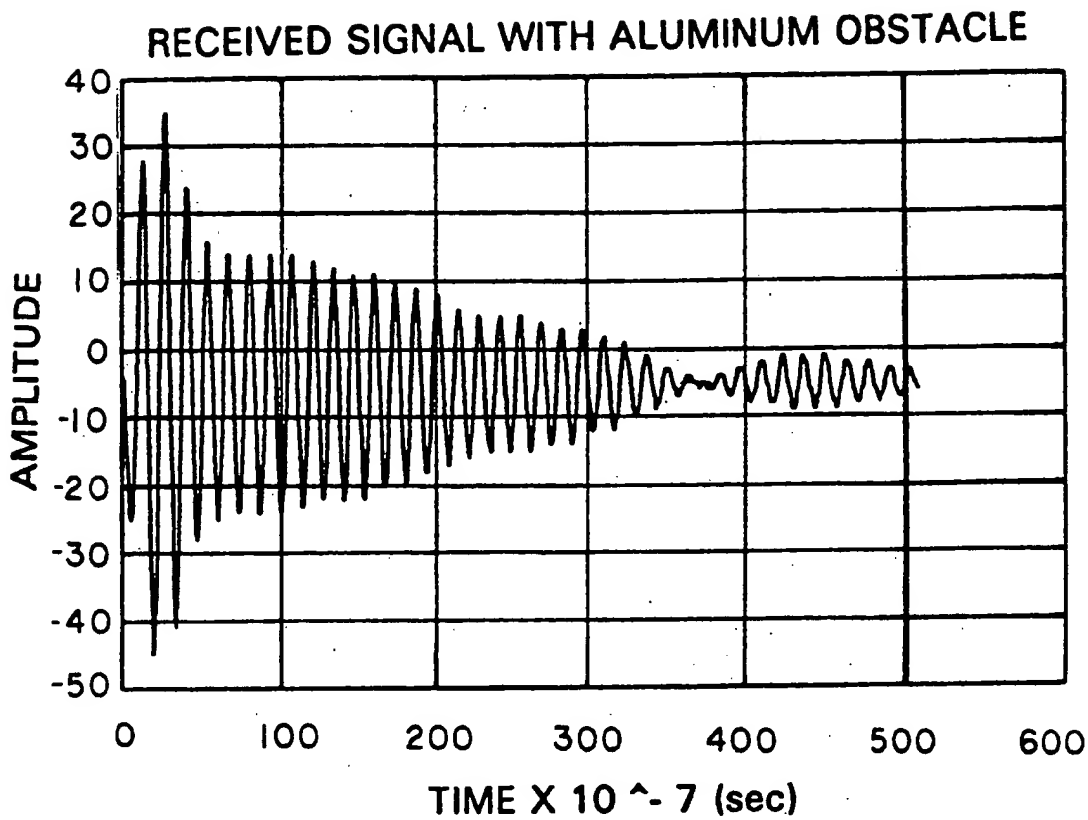
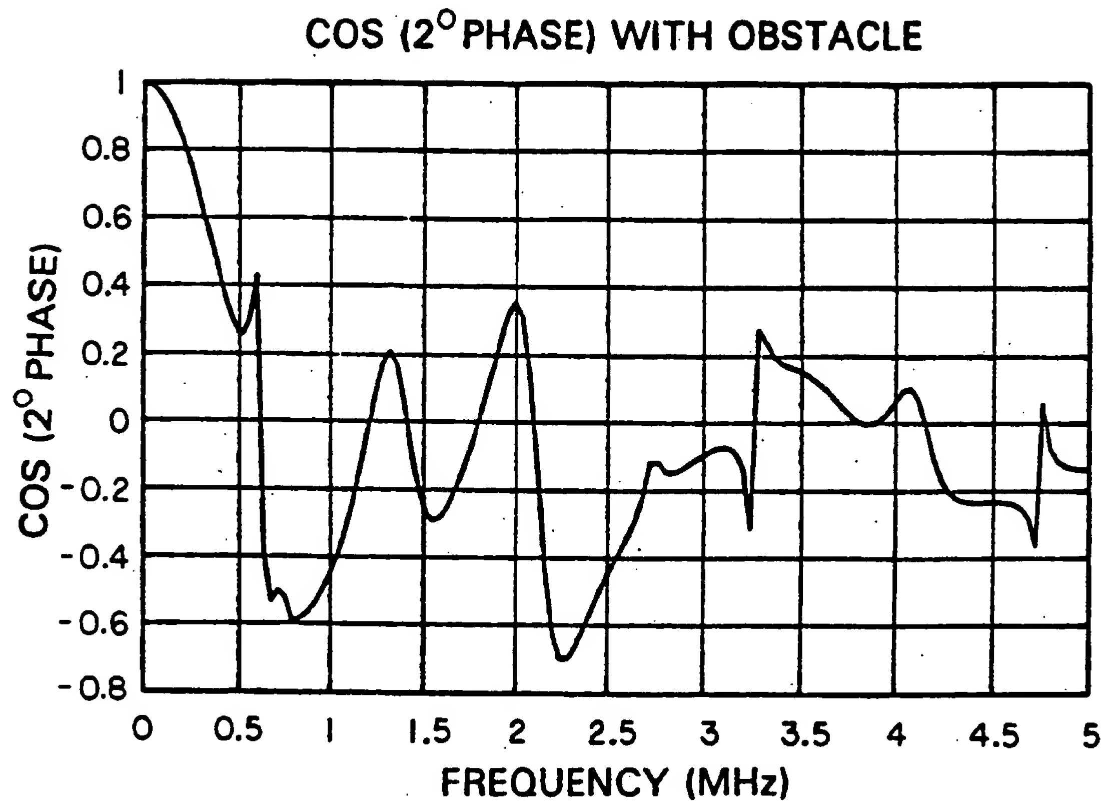

RECEIVED SIGNAL WITH ALUMINUM OBSTACLE
40
30
20
10
0
-10
-20
-30
-40
-50
AMPLITUDE
0
100
200
300
400
500
600
TIME X 10 ^- 7 (sec)

FIG.-8



COS (2° PHASE) WITH OBSTACLE
1
0.8
0.6
0.4
0.2
0
-0.2
-0.4
-0.6
-0.8
COS (2° PHASE)
0
0.5
1
1.5
2
2.5
3
3.5
4
4.5
5
FREQUENCY (MHz)

FIG.—9

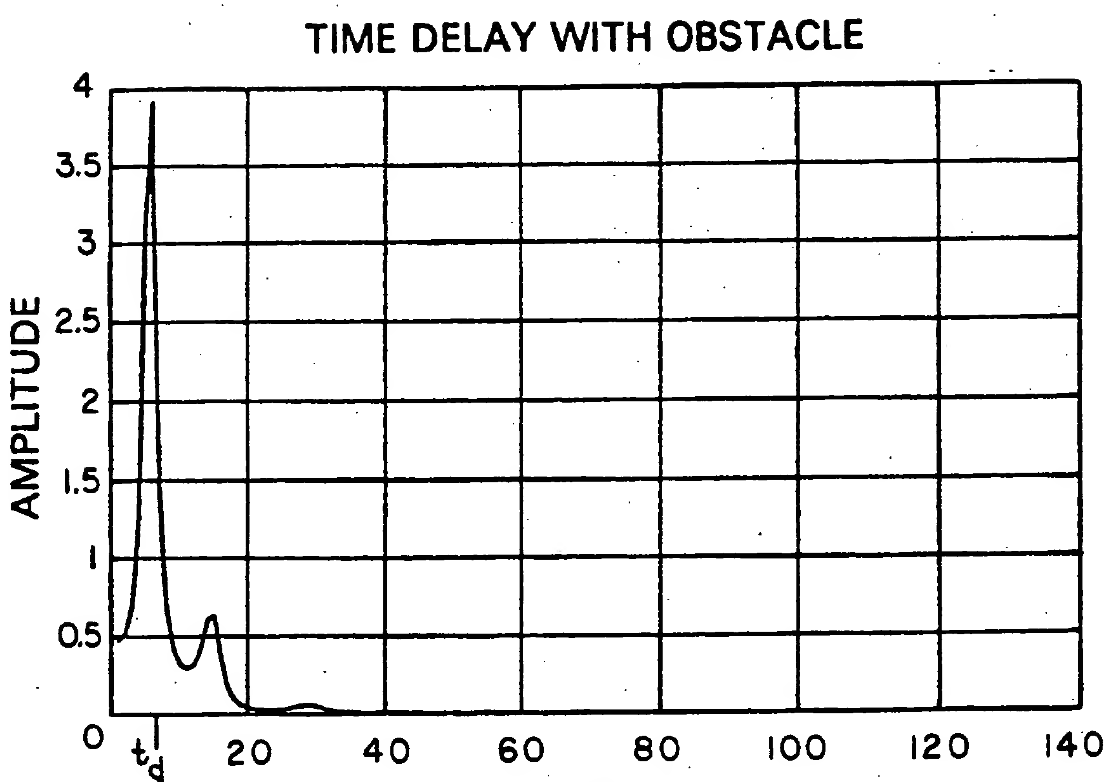

TIME DELAY WITH OBSTACLE
AMPLITUDE
4
3.5
3
2.5
2
1.5
1
0.5
0
t_d
20
40
60
80
100
120
140
TIME X 10 ^- 7 (sec)

FIG.—10

METHOD AND APPARATUS GENERATING HIGH RESOLUTION DATA AND ECHO IDENTIFICATION

TECHNICAL FIELD

The present invention is generally directed to on apparatus and method of echo identification wherein the sources of echoes generated from interface boundaries in a medium may be more effectively identified in the processing of obtained waveform data to generate higher resolution in the obtained data. The method of the present invention is particularly useful for performing ultrasound reflectometry and medical sonography by obtaining data representative of a medium being studied and suppressing signals from overlapping echo waveforms generated from obstacles adjacent on object under test. More particularly, the method of the invention enables the sources of such echo waveforms to be more effectively identified to yield an indication of the contribution of overlapping echoes to generate higher resolution data signals.

BACKGROUND OF THE INVENTION

In various disciplines such as ultrasound reflectometry, medical imaging using sonography, radar, seismology, echocardiography and other similar areas, a common problem exists in that waveform data obtained in these areas will often include reflections from adjacent obstacles which will often overlap the desired echo or reflected signals. The overlapping echo waveforms tend to obscure the desired signals and result in lower resolution which inhibits effective and accurate interpretation of the data. I ultrasound reflectometry which is used for medical imaging as well as non-destructive testing applications, such obscuring echoes result in the inability to form high resolution images thereby degrading the diagnostic capabilities achievable by the procedure. Medical imaging has become an important aspect of present day medical technology and presently includes techniques such as X-ray tomography, magnetic resonance imaging (MRI) or CAT-scan techniques which themselves require relatively high cost, elaborate equipment and processing capabilities which inhibit their effective use in many situations. If higher resolution and improved ultrasound images could be obtained, this technique may considerably reduce the cost as compared to other imaging techniques while providing more accurate diagnostics in medicine.

Ultrasound has been known to be applicable for medical imaging as well as non-destructive testing as it is a mechanical wave phenomena wherein a medium under test will enable generated ultrasonic pulses or waves to propagate therein. In medical imaging techniques such a medium will be the soft tissue of the body with reflecting objects being the internal organs thereof. The waveforms used in ultrasound reflectometry as well as medical sonography are typically several cycles in length and are rapidly damped to a few wavelengths, at the resonant frequency of the transducer. The transducer is typically a piezoelectric crystal which gives rise to a sonic wave propagating into the medium under study and reflecting as echoes from reflecting surfaces within the medium which are detected by the reverse of the piezoelectric effect. In medical sonography for example, the transducer normally serves as both a transmitter and a receiver, and therefore induced vibrations must be damped quickly to avoid a very long ring-down time in order to receive echo signals. When low frequencies are chosen to obtain low attenuation of the waveforms through the medium under test, then the time waveforms of reflections from adjacent obstacles will often overlap. The overlap of echoes makes it difficult to interpret where the sources of the echoes are located and results in extremely difficult signal processing problems to obtain high resolution data which avoids the contribution of unwanted overlapping echo information.

One processing technique being echo envelope processing has been utilized, but it has been found that the received information is still significantly contaminated by obscuring overlapping echoes. The overlapping echo waveform phenomena resulting from the low frequencies chosen for ultrasound reflectometry as based upon attenuation and resolution constraints have been noted in literature such as found in an article by R. C. Kemerait and D. G. Childers entitled "Signal Detection and Extraction By Cepstrum Techniques" IEEE Transactions on Information Theory, Volume II-18, No. 6, pp. 745–749, November, 1972, as well as a publication by J. Blitz, entitled "Ultrasonics: Methods and Applications", Van Nostrand Reinhold Company, New York, 1971. In the first of these publications, there is set forth a technique for decomposing a composite signal of unknown multiple wavelets which overlap in time. Several prior procedures for achieving the decomposition of superimposed signals include inverse filtering wherein a signal is transformed by a linear time-invariant system, whose Fourier transform is the reciprocal of the transform of the signal components to be removed. In such a method, the signal must be known and the signal to noise ratio must be quite large. Decision Theory has also been used to decompose superimposed signals to estimate the echo amplitude and arrival times, but only if the signal wave shape is known. If the wavelet waveshape and number of echoes are unknown, other techniques have been looked to. The Cepstrum techniques of echo detection accomplish decomposition by means of a function of the power spectrum of the received signals to determine the timing and relative amplitudes of echoes in the system. The signal waveform is then extracted by means of complex Cepstrum techniques.

It has also been found that phase information can be used in the estimation of time delays between two received signals as described in an article by A. G. Piersol, entitled "Time Delay Estimation Using Phase Data", IEEE Transactions on Acoustics, Speech and Signal Processing, Volume ASSP-29, No. 3, pp. 471–477, June 1981. Such an estimation of time delays between two received signals using phase measurements relied upon the use of straightforward regression analysis procedures on phase estimates at properly selected frequencies in the frequency domain. Such analysis of the phase information was found to yield time delay estimates having realistic error assessments based upon non-parametric variance calculations.

Similarly, a publication by T. F. Quatieri, Jr. and A. B. Oppenheim, entitled "Iterative Techniques for a Minimum Phase Signal Reconstruction from Phase or Magnitude", IEEE Transactions on Acoustics, Speech and Signal Processing, Volume ASSP-29, No. 6, pp. 1187–1193, December, 1981, develops iterative algorithms for reconstructing a signal from a partial specification thereof in the time or frequency domains. The

technique utilizes iterative algorithms for reconstructing a minimum or maximum phase signal from the phase or magnitude of its Fourier transform. Additionally, a phase unwrapping algorithm is proposed which is implemented by applying the Hilbert transform to the logarithmic value of the Fourier transform of a minimum phase sequence.

Although the information content of the phase component of systems has been recognized in different applications, the phase unwrapping techniques have inhibited their effective use with respect to processing in ultrasound reflectometry or similar systems where obscuring overlapping echoes degrade resolution. For example, in a publication by E. Poggiagliomi, A. J. Berkhout, M. M. Boone, entitled "Phase Unwrapping, Possibilities and Limitations", Geophysical Prospecting, No. 30, pp. 281–291, 1982, a phase unwrapping technique is set forth wherein the phase spectrum can only be correctly unwrapped between notches in the amplitude spectrum and other experimental difficulties are also present in the described method.

Some researchers have turned to the principle of Maximum Entropy, which involves autoregressive modeling, and which has been applied successfully to generate power spectral estimations. In publication by J. P. Burg, entitled "Maximum Entropy Spectral Analysis", Modern Spectrum Analysis, edited by D. G. Childers, IEEE Press, New York, pp. 34–41, 1978. Comparison of the maximum entropy method to some traditional techniques was analyzed in a book by A. V. Oppenheim and R. W. Schafer, entitled "Digital Signal Processing, Printice-Hall, Englewood Cliffs, New Jersey, 1975, wherein the maximum entropy method was found to produce higher frequency resolution and yield data of better dependability as it depends only upon the available data and requires simplified storage requirements owing to the infinite impulse response structure of the resulting all pole filter. The all pole filter is found to yield a smoothly changing phase estimate, which does not appear to have the experimental difficulties reported with various other methods to obtain phase estimates. Such a method has been studied by N. Erdol in a PhD dissertation entitled "Use of the Maximum Entropy Method for Phase Estimation" at the University of Akron, Akron, Ohio. In this study, it was found that phase information obtained by using maximum entropy spectral estimates could be utilized to subsequently obtain delay information from the maximum entropy phase estimation by means of discrete Fourier transformation techniques.

SUMMARY OF THE INVENTION

Based upon the foregoing, it has been found that there is a need to provide an apparatus and method which can be used to identify the source of echo signals to enable higher resolution data to be obtained by minimizing the contribution of overlapping echoes which may obscure the desired echo signals. The present invention is directed to providing a method and apparatus for identifying the locations of the sources of echoes contributing to data received in systems in applications such as ultrasound reflectometry, medical sonography, radar/sonar applications, acoustic non-destructive evaluation and testing, seismology, echocardiography and the like. The present invention is also directed to a method and apparatus for improving the resolution of received linear frequency modulated signals found in such applications, and particularly to a method and apparatus for performing ultrasound reflectometry and identifying the source of echoes in this application.

Generally, the method comprises the steps of obtaining echo signals in the form of analog data and thereafter transforming the analog data signals into a digital data sequence. A phase estimate of the digital data sequence is obtained by generating a power spectral estimate of the data signals by means of a technique known as the Maximum Entropy estimation method wherein the power spectral estimate is processed to yield the phase information. Using phase information, an estimate of the delay time of received echo signals which are reflected from an interface boundary in a medium can be obtained. With an estimate of the delay time, the distance to the interface boundary may be determined to locate the sources of echo signals, wherein such information may then be utilized to minimize the contribution of overlapping echoes to enhance the resolution of the desired echo signals in the system.

It is therefore a main object of the invention to provide a method for echo identification to improve the resolution of received echo signals.

It is another object of the invention to provide for high resolution of small structures while retaining low attenuation of current signal sources.

Another object of the invention is to provide an apparatus and method whereby phase information may be obtained from a digital data sequence of reflected echo signals to yield an estimate of delay time which can be far less than the reverberation time of obscuring echoes to minimize reverberation artifacts.

Yet another object of the invention is to provide a method for improving the resolution of linear frequency modulated signals by using the method of Maximum Entropy at all computational stages which relies only upon available data and provides higher frequency resolution as well as simple storage requirements owing to the infinite impulse response structure of the resulting all pole filter utilized in the method.

BRIEF DESCRIPTION OF THE DRAWINGS

These and other objects of the invention will become more readily apparent from a reading of the detailed description of the invention in conjunction with the drawings, wherein:

FIG. 1 shows an apparatus for generating and receiving ultrasound through an experimental medium, typifying the apparatus of the invention as used in ultrasound reflectometry;

FIG. 2 sets forth a simplified representation of an ultrasonic pulse signal path within a test medium indicating the sources of echoes resulting from interface boundaries within the medium or obstacle, which may result in overlapping echo information;

FIG. 3 sets forth a block diagram representation of the preferred method for echo identification and improving resolution as associated with a technique such as ultrasound reflectometry;

FIG. 4 sets forth a block diagram setting forth in more detail the steps of the method in the present invention to identify the sources of echo information;

FIGS. 5–7 set forth graphical representations of experimental results showing a received echo signal without multiple reflections wherein the amplitude will provide digitized values for subsequent processing, the phase information obtained therefrom by the method of the invention and the estimate of the time delay of the received signals respectively; and

FIGS. 8–10 set forth graphical representations of received echo signals with multiple reflections present yielding phase information by means of the method in the present invention and an estimate of the time delay respectively giving an indication of the source of echo information to enable estimation of echo contributions and to improve resolution of the received signals.

DETAILED DESCRIPTION OF THE INVENTION

Turning now to FIG. 1, an apparatus usable to obtain data as well as to perform processing on the received data is shown for use in an ultrasound reflectometry method. The invention will be described in terms of ultrasound reflectometry but it should be understood that the apparatus and method may be modified to suit other applications such as radar, seismology, echocardiography and other similar technologies. The apparatus as shown in FIG. 1, shows an experimental arrangement to show the method of the invention for a simplified arrangement comprising a water tank **10** having a simple reflecting structure **12** therein. The experimental arrangement including tank **10** includes a first ultrasonic pulse producing transducer **14** arranged on one side of the reflecting structure **12** and a second ultrasonic transducer **16** arranged on the opposite side of the reflecting structure **12** to receive echo waveforms. It should be recognized that depending upon the particular application of the invention, one or more transducers may be utilized. For example, in the application of medical sonography, one ultrasonic transducer will both generate ultrasonic pulses and detect echoes that return from reflecting surfaces within the body being studied. Typically, an ultrasound transducer comprises a thin piezoelectric crystal, which is made up of dipoles that are firmly bound in a crystalline structure. In ultrasound reflectometry including medical sonography, the piezoelectric crystal is made to vibrate at its natural frequency by a single voltage pulse generated by an output pulse circuit **18** having power supply **20** and a triggered output timing circuit **22**. The waveforms conventionally used in ultrasound reflectometry are normally several cycles in length, wherein a typical pulse length is between 1 to 10 microseconds. Low frequencies are normally chosen to obtain low attenuation through the medium under test and the output pulse is rapidly damped to maintain a short ring-down time between the initiation of the ultrasonic wave and the cessation of all vibrations in the transducer. The ultrasound waveforms generated by transducer **14** will be detected by transducer **16** in the form of analog data signals which are transformed by a conventional A/D converter to enable display on a digital oscilloscope **24** after which the digital data signal may be processed by the method of the invention using computer **26**. It is again reiterated that the particular experimental apparatus shown in FIG. 1 may be modified in a known manner by those of ordinary skill for use of the invention in the other areas of technology as previously described, wherein linear frequency modulated signals generated from echo waveform data may be obtained by suitable detection means and subsequently processed to yield an estimate of the echo contributions and to identify the source of such echoes to yield high resolution data or images.

In FIG. 2, there is shown the signal path for an incident wave through an obstacle in its path giving some idea of echo waveform contributions for a simple reflecting structure in the path of an incident wave. When an ultrasonic beam is incident upon a boundary between two interfaces, part of the beam will be reflected from the boundary as shown at **30** while part of the beam is transmitted into the second medium shown at. **32** being the obstacle of FIG. 2. Within the obstacle, a second interface is encountered where again part of the incident beam on the second boundary will be reflected as shown at **34** while part will be transmitted through the boundary into the medium lying beyond the boundary as shown at **36**. This process continues within the obstacle creating reverberations wherein the ultrasonic beam will be attenuated by absorption and scattering until it is ultimately dissipated in the form of random molecular motion or heat. As shown in FIG. 2, an incident wave upon an obstacle having two boundary surfaces will create a number of reverberation echo signals wherein each of the echoes will have a different delay time corresponding to the thickness of the obstacle as shown in FIG. 2. The fraction of the incident beam intensity which is reflected depends upon the acoustic impedances of the mediums about the boundary, and thus the particular medium being tested will directly influence the amount of reverberation occurring within the system and the potential of generating additional echo waveforms in the system. It is also noted that the fraction of the ultrasonic beam that remains after reflection, which is transmitted into the second medium, will in general deviate from its original direction wherein the degree of deviation will depend upon the relative velocity of sound in the two media. The direction of the transmitted beam can be obtained by the application of Snell's law similar to light.

Although a relatively simple reflecting structure is shown in the apparatus of FIG. 1, in medical imaging the medium through which the ultrasonic pulses will travel will be the soft tissue and the reflecting objects will be the internal organs. Due to the low frequencies used in ultrasound reflectometry to obtain low attenuation through the medium under test, the time waveforms of reflections from adjacent obstacles, particularly in more complex systems, will often overlap. The overlap of echoes makes it difficult to interpret where the sources of the echoes are located, and thus to account for their contributions in the received echo signals thereby tending to degrade resolution of the received signals.

Turning to FIG. 3, there is provided a block diagram of the apparatus and method in the invention to identify the source of echo information. In general, the apparatus and method of the invention include means for detecting echo waveform data to be digitized thereby forming a digital data sequence as indicated at **50**. The digital data signal is thereafter processed at **52** to obtain a phase estimate of the data signal containing information correlating to the time delay of the echo waveforms detected in the system. At **54**, means are provided to obtain an estimate of the delay time of the echo waveform data from the phase estimate obtained at **52**. Finally, at **56** there are means to obtain a distance to a reflecting boundary being the source of the echo waveform data which correlates to the delay time calculated in step **54**. These steps form a very generalized description of the method of the invention as a means of echo identification of received waveform data.

The invention will now be described in more detail relative to the general procedures as outlined with reference to FIG. 3. Turning to FIG. 4, the method of the invention includes a first step in computing the coeffici-

ents of a minimum phase filter H(z) to form a prediction filter

$$Y(z) = \frac{\sigma}{H(z)} \quad (1)$$

which adequately represents the digital data signal x(n). Prediction filters are computed at **62** for the even and odd data sequences generated at **60**. The design of the prediction filter, represented by the even and odd constituents of the data sequence x(n), is particularly important in applications such as ultrasound reflectometry and other similar environments where the digital data signal x(n) is only partially available due to the nature of the data and its method of acquisition. In such systems, the unknown portion of x(n) must be predicted and the power spectral density is necessarily estimated, which require the predictive filter to yield as accurate a prediction of the unknown portion of the data as possible. As the power spectral density can be computed only if all of the data x(n) is known, the two methods including the prediction of the unknown portion of x(n) and the estimation of the power spectral density are related and will benefit from a model which uses only the available data to accomplish this task.

Modeling the system as an all pole, minimum phase filter provides a linear model, but it is noted that the processing preformed in the method is not linear. The all pole, minimum phase filter also simplifies storage requirements owing to the infinite impulse response structure of the filter.

The unit sample response y(n) of Y(z) is not directly related to the data x(n), however it can be shown that their respective autocorrelation values are related by:

$$R_{xx}(n) = R_{yy}(n) n = 0, 1, \ldots, N \quad (2)$$

wherein N is the order of the all pole, minimum phase filter. It thus can be seen that $R_{yy}(n)$, for n greater than N, provides a non-zero extension to $R_{xx}(n)$. Since the power spectrum is the Fourier transform of its autocorrelation sequence, it can thus be shown that

$$|Y(e^{j\omega})|^2 = S_{yy}(e^{j\omega}) = \left| \frac{\sigma^2}{H(z)H(1/z)} \right|_{z=e^{j\omega}} \quad (3)$$

is an estimate of the power spectrum of the data x(n). The estimation of the power spectrum is independent of the phase of $x(e^{|j\omega|})$, and the error noise power of the predictive filter is σ^2.

The invention utilizes the above mentioned filter design methodology to determine the distance to a source of echo waveform data such as the distance to a tissue interface in a medical sonography application. A brief description of the mathematical model to determine this distance is shown in the following.

A wide-band, short time duration pulse x(t), and its reflections x from an interface boundary at a distance d_i can be represented at a receiver as:

$$x_i(t) = x(t) + a_1 x(t-t_1) + a_2 x(t-t_2) + \quad (4)$$

where the delay time of the waveform data from the interface boundary, $t_d = 2d_i/v$, being the round trip distance to the interface boundary divided by the speed of the reflections in the medium. The term, x, represents the widening of the echo waveform as it travels through the medium and a_i accounts for the attenuation of the waveform in the medium. As mentioned before, short distances involved and the widening of the echo waveform result in echoes which overlap. Windowing techniques are used to remove the overlapping parts of the received waveforms, but are often found to severely truncate the reflected waveform. The signal to be treated after windowing can be represented as:

$$y(t-t_1-\epsilon) = a_1 x(t-t_1) w_\tau(t-t_1-\epsilon) \quad (5)$$

where τ is the width of window ,w, being symmetrical about $t_1+\epsilon$, which is expected to be the center of data of the reflected signal. Due to the effects of dispersion in the medium, the center of the reflected signal is not as well defined as the incident signal, and it is thus the goal to compute $t_1 = 2d_1/v$ to enable computation of the distance d_1.

It is known that the incident signal x(t) is symmetrical and it is reasonable to model the medium through which the echo waveform travels as a linear phase system. Translating the center of the window to the origin (t=0) causes the center of the reflected signal to be ϵ away from the origin, where e may be positive or negative. Since a translation by ϵ in time implies a linear phase term in the frequency domain, the problem becomes one of finding the linear phase component of $a_1 x(t-\epsilon)$ from its asymmetrically truncated form $y(t-\epsilon)$.

The Maximum Entropy spectral estimation philosophy and methodology is utilized to estimate the phase and thus the delay time and therefore the distance to the interface boundary being a source of echoes. The Maximum Entropy Method enables finding an accurate estimation of the phase of a signal which would normally require complete knowledge of the signal, but wherein such knowledge is unavailable due to windowing. Although the Maximum Entropy Method uses a non-linear criterion of goodness, the predictive filter which is computed is a linear filter which generates a substitute signal for the truncated one caused from windowing. This new signal is not truncated but is not a point by point equivalent of the truncated one, although providing an accurate estimation. The power spectrum of the substitute signal has been found to be very close to the power spectrum of the original signal before truncation, and thus addresses the problem of using a power spectral estimation method to obtain phase estimation.

For simplicity of notation, we will use x(t) to denote the signal whose phase we wish to estimate. The signal x(t) is sampled to give a number of equally spaced sampled data points x(n) whose Fourier transform is $X(e^{j\omega})$. At 66 in FIG. 4, there are provided means to obtain the phase estimate of the sample data for the process x from the known data samples x(0), x(1), . . . , x(M), where M is the data length being an even integer. To obtain the phase estimate, the time series x(k) is written as the sum of an even series, e(k), and an odd series o(k), where the point of symmetry is chosen at $N_S = M/2$. The power spectra of the time series and its even and odd components are related by

$$S_{xx}(e^{j\omega}) = S_{oo}(e^{j\omega}) + S_{oo}(e^{j\omega}) \quad (6)$$

wherein the Fourier transforms of each may be represented by

$$X(e^{j\omega}), E(e^{j\omega})e^{-j\omega T}, jO(e^{j\omega})e^{-j\omega T}, \quad (7)$$

with the even and odd components having real and imaginary portions. The phase $\theta(\omega)$ may be defined as

$$\theta(\omega) = \text{Arctan}\,\frac{O(e^{j\omega})}{E(e^{j\omega})} \quad (8)$$

where $O(e^{j\omega})$ and $E(e^{j\omega})$ are the real functions of ω for the even and odd components and the factor $e^{-j\omega T_s}$ is due to the fact that the origin of symmetry is $T_s = N_s\Delta t_o$ where N_s is the number of data samples to the center of the data block and Δt_o is the sample spacing. From this, it may be shown that

$$\text{Cos}[2\theta(\omega)] = \frac{S_{oo}(e^{j\omega}) - S_{oe}(e^{j\omega})}{S_{ee}(e^{j\omega}) + S_{oe}(e^{j\omega})} \quad (9)$$

wherein the even and odd power spectral terms $S_{ee}(e^{j\omega})$ and $S_{oo}(e^{j\omega})$ may be expressed in terms of the even and odd components of the predictor polynomial H(z) yielding

$$\text{Cos}[2\theta(\omega)] = \frac{P_oH_o(e^{j\omega})H_o(e^{-j\omega}) - P_oH_o(e^{j\omega})H_o(e^{-j\omega})}{P_oH_o(e^{j\omega})H_o(e^{-j\omega}) + P_oH_o(e^{j\omega})H_o(e^{-j\omega})} \quad (10)$$

To be consistent with the principles of Maximum Entropy, the point of even and odd symmetry is chosen at the median of the available data at $N_s = M/2$. The even series, e(k) and the odd series o(k), are put through their respective Maximum Entropy filters wherein the outputs will be the power spectra of e_k, o_k accordingly. To obtain the cosine of twice the phase of x_k, the ratio of the difference and the sum of the even and odd series power spectra are computed.

Since the Maximum Entropy filters through which the even and odd series are processed are all pole, minimum phase filters, the right hand side of Equation 10 converges for all values of ω, and the phase $\theta(\omega)$ can be calculated from Equation 10.

After the phase estimate of the sample data has been obtained by the Maximum Entropy Method, this estimate is used at 68 which provides a means to calculate the time delay of received echo waveforms by obtaining the power spectrum of the phase estimate, again using the Maximum Entropy Method. As an example, a pulse of ultrasonic energy traveling through a uniform lossless medium will be reflected from a boundary and the echo will be received at the point of origin of the signal. The incidence signal, x(t) of ultrasonic energy will travel a total distance 2d through the medium which has a velocity of propagation v. The received signal, y(t), will be a function of the delay time, t_d, where $t_d = 2d/v$. By preprocessing the echo signal according to the Maximum Entropy Method to yield the phase estimate described at 66, it may be shown that

$$\text{Cos}2\theta(\omega) = \text{Cos}[2\omega t_d] \quad (11)$$

where the argument will be proportional to the delay time of the echo with all data referenced to the origin of symmetry being the median of the available data acquired from the reflected echo waveform. The amount of delay with respect to a known point, being the point of even and odd symmetry, is captured as the "frequency" of a sinusoid. The "frequency" of the sinusoid can be extracted by obtaining the Maximum Entropy power spectrum of the phase estimate of the original data or COS28. The term "frequency" has been embedded in quotation marks because it has the unusual dimension of time incorporated therein.

Once the delay t_d is known, the distance d to the boundary may be determined at 70 which provides means to estimate echo contributions by determining the source of the echo waveform data.

Turning now to the experimental results shown in the remaining Figures, the apparatus and method of the invention were applied to a simplified experimental model to give some indication of the improved resolution and method of identifying the location of the sources of echoes in the sampled data. In the experimental model, a sheet of aluminum having a thickness 0.16 cm was placed into the path of ultrasound similar to that shown in FIG. 1. In the apparatus of FIG. 1, the ultrasound transducers **14** and **16** were 3.5 MHz focused transducers, mounted on opposing sides of a plastic tank containing water. In the analog to digital conversion of the received data, the data were oversampled at 20 M samples/second and then averaged for an effective sample rate of 10 M samples/second. Although the transducer energy is centered around the 3.5 MHz resonance of the transducer, the information content in the signal phase and amplitude is distributed around the unit circle in the z plane. It has been found that the poles of the all pole predictive filter convey their information as the unit circle is traversed, wherein the most rapid changes occur in the vicinity of the poles. Although the signal energy will be concentrated near the 3.5 MHz transducer frequency, the computation of the prediction error filter coefficients, and therefore the pole locations, will evidence the information from the poles of the predictive filter on the whole unit circle. If only the information on the unit circle adjacent to the poles would be to window the result, thereby eliminating useful information for the Maximum Entropy phase estimation. Therefore, the entire unit circle is traversed in the Z plane using transform calculations to obtain the entire information content of the waveform data for the phase estimation using the Maximum Entropy Method.

Further in the experimental configuration, a bi-junction transistor was used to discharge a capacitor across the transmitter transducer **14** to obtain a sufficiently large signal. The received signal at transducer **16** was sufficiently large to be digitized by a Tektronix Model 2220 digital oscilloscope at **24** without prior amplification.

In a first example, a direct transmission of the pulse from the 3.5 MHz transducer **14** to receiving transducer **16** of the same frequency is shown, wherein the obstacle **12.5** removed from the water tank **10** in the apparatus of FIG. 1. Turning to FIG. 5, the signal without an obstacle in the ultrasound path is shown which does not have multiple reflections therein. The amplitude of the received waveform is the digitized value, which may be regarded as a relative value. Utilizing the method of the invention as previously described, the phase estimate $\theta(\omega)$ can be calculated from Equation 10 which gives the cosine of twice the phase angle being shown in FIG. 6. Proceeding with the method as previously defined, the spectrum of the phase estimation data as shown in FIG. 6 yields the echo identification giving a delay time as shown in FIG. 7.

Turning to FIG. 8, after the thin aluminum sheet is placed in the water of tank **10** as an obstacle shown at **12** in FIG. 1, the received signal shows a long, ill-defined reflected signal indicating multiple reflections as con-

trasted with FIG. 5. The extended echo of FIG. 8 appears without clear delineation between the different components of the multiple echo, making interpretation of the obtained waveform data extremely difficult. Using the method of the invention, the phase estimation being the cosine of twice the phase angle is shown in FIG. 9, wherein the oscillatory nature of the curve enables identification of the multiple echoes as shown in FIG. 10. In FIG. 10, the position of the first peak corresponds to the time of transit through the aluminum sheet which comprises the test obstacle. This transit time is the round trip or twice the path length within the 0.16 cm obstacle. The distance, d, to the boundary may be computed from the time delay calculation as shown in FIG. 10, wherein for the experiment described above, this sample calculation is

$$d=\tfrac{1}{2}(6420m/s)\,(0.5us)=1.6mm$$

which is seen to verify and identify this source of multiple echoes. It is also noted that the relative amplitudes of the multiple echoes are consistent. Based upon this simplified experimental arrangement, the computed value of t_d was found to be quite accurate and only limited by the resolution obtained in digital transformation of the data. The resolution can be improved by increasing the number of samples, but at the expense of increased filter lengths in the processing. Further experimental results show that the method provides stable delay detection with respect to variation of data length and additive noise, with accuracy being a function of filter length only. Although a simplified model has been used to give an indication of results obtainable by the method and apparatus of the invention, it should be apparent that the invention can be extended to more complex systems in a variety of areas such as medical sonography, ultrasound non-destructive testing, echocardiography, underwater sonar, radar and geophysical data as previously described. The invention may also be applied to dispersive media to detect group delays as well as phase delays beneficially.

Although the invention has been described with respect to ultrasound reflectometry in some detailed aspects thereof, it should be obvious that many modifications and variations of the present invention are possible in the light of the general concepts of the disclosed invention. It is therefore understood that the invention may be practiced other than as specifically described herein and is only limited by the scope of the appended claims.

What is claimed is:

1. A method for performing waveform processing comprising the steps of:

obtaining waveform signals in the form of analog data signals by means of detection means,

transforming said analog data signals to a digital data signal,

obtaining a phase estimate of said digital data signal by generating a power spectral estimate of said digital data signal by means of a technique known as the Maximum Entropy Estimation Method, and processing said power spectral estimate to yield said phase estimate,

obtaining an estimate of the delay time of said waveform signals which are reflected from an interface boundary in a medium,

determining the distance to said interface boundary by means of said estimate of the delay time to locate the sources of said waveform signals,

processing said waveform signals to account for the sources generating these signals to minimize the contribution of predetermined waveform signals.

2. A method for performing waveform processing as in claim 1, wherein,

the step of obtaining said phase estimate of said digital data signal includes modeling the system as an all pole, minimum phase filter which yields a smoothly changing phase estimate of said digital data signal.

3. A method for performing waveform processing as in claim 1, wherein,

said step of obtaining a phase estimate of said digital data signal includes the steps of utilizing the even and odd components of said digital data signal and computing a prediction filter for each of said even and odd components from which said power spectral estimate can be generated.

4. A method for performing waveform processing as in claim 1, wherein,

said step of obtaining a phase estimate of said digital data signal includes computing the coefficients of a minimum phase filter to generate a prediction filter which adequately represents said digital data signal in systems where said digital data signal is only partially available and the unknown portion of said digital signal must be predicted.

5. A method of performing waveform processing as in claim 1, wherein,

said step of obtaining a phase estimate includes generating a power spectral estimate of the even and odd components of said digital data signal wherein the phase, $\theta(\omega)$, is related to the power spectral estimate as represented by the following:

$$\mathrm{Cos}[2\theta(\omega)]=\frac{S_{oo}(e^{j\omega})-S_{oe}(e^{j\omega})}{S_{ee}(e^{j\omega})+S_{oe}(e^{j\omega})}$$

which may be shown to yield

$$\mathrm{Cos}[2\theta(\omega)]=\frac{P_oH_o(e^{j\omega})H_o(e^{-j\omega})-P_oH_o(e^{j\omega})H_o(e^{-j\omega})}{P_oH_o(e^{j\omega})H_o(e^{-j\omega})+P_oH_o(e^{j\omega})H_o(e^{-j\omega})}$$

wherein the phase $\theta(\omega)$ can then be calculated.

6. A method for performing waveform processing as in claim 5, wherein,

said step of obtaining an estimate of the delay time utilizes the phase estimate which is related to the delay time as follows:

$$\mathrm{Cos}2\theta(\omega)=\mathrm{Cos}(2\omega t_d)$$

where the argument is proportional to the delay time of the echo waveform data.

7. A method for performing waveform processing as in claim 6, wherein,

said step of determining the distance to said interface boundary utilizes the estimate of the delay time wherein the distances related to the delay time by the following:

$$t_d=2d/v$$

where d is the distance to said interface boundary and v is the velocity of the waveform signal in the medium through which it travels.

8. A method for performing waveform processing as in claim 1, wherein,

said waveform signals are echo signals and said step of determining the distance to said interface boundary enables identifying the echo location of the source of echoes wherein said waveform data includes overlapping echoes which tend to obscure waveform data, wherein identification of the echo location of the source of said overlapping echos allows minimization of the contribution of said overlapping echos.

9. A method for performing waveform processing as in claim 1, wherein,

said step of obtaining an estimate of the delay time includes obtaining a power spectrum of the phase estimate by means of the Maximum Entropy Estimation Method, wherein the power spectrum of the phase estimate curve exhibits a peak at a time corresponding to said delay time.

10. A method for performing waveform processing as in claim 1, wherein,

said waveform data is generated in an ultrasound reflectometry system.

11. A method for performing waveform processing as in claim 1, wherein,

said waveform data is generated in a medical sonography application.

12. A method for performing waveform processing as in claim 1, wherein,

said waveform data is generated in an acoustic non-destructive evaluation and testing application.

13. A method for performing waveform processing as in claim 1, wherein,

said waveform data is generated in a radar system.

14. A method for performing waveform processing as in claim 1, wherein,

said waveform data is generated in a sonar system.

15. A method for performing waveform processing as in claim 1, wherein,

said waveform data is generated in a seismology application.

16. A method for performing waveform processing as in claim 1, wherein,

said waveform data is generated in an echocardiography application.

17. A method for performing ultrasound reflectometry comprising the steps of:

generating ultrasound waveforms within a medium by a transducer means,

detecting ultrasonic echo signals in the form of analog data signals by transducer means,

transforming said analog data signals to a digital data signal,

obtaining a phase estimate of said digital data signal by generating a power spectral estimate of said digital data signal by means of a technique known as the maximum entropy estimation method, and processing said power spectral estimate to yield said phase estimate,

obtaining an estimate of the delay time of said ultrasound echo signals which are reflected from an interface boundary within said medium,

determining the distance to said interface boundary by means of said estimate of the delay time to locate the sources of said echo signals,

processing said echo signals to account for the sources generating the ultrasound echo signals wherein the contribution of predetermined echo signals is minimized.

18. A method for performing ultrasound reflectometry as in claim 17, wherein,

the echo waveform data is generated in a medical sonography technique.

19. A method for performing ultrasound reflectometry as in claim 17, wherein,

the echo waveform data is generated in a acoustic non-destructive evaluation and testing application.

20. A method of identifying the echo location of the source of echoes in echo waveform data, which comprises the steps of:

obtaining echo waveform signals in the form of analog data signals by means of detection means,

transforming said analog data signals to a digital data signal,

obtaining a phase estimate of said digital data signal by generating a power spectral estimate of said digital data signal by means of a technique known as the Maximum Entropy Estimation Method, and processing said power spectral estimate to yield said phase estimate,

obtaining an estimate of the delay time of said echo signals which are reflected from an interface boundary in a medium,

determining the distance to said interface boundary by means of said estimate of the delay time to locate the sources of said echo signals.

21. An apparatus to perform waveform processing, comprising:

means to generate waveforms in a medium which will be reflected for interface boundaries within said medium to form said echo waveforms in said medium,

means to detect said echo waveforms to generate analog echo data signals,

means to transform said analog echo data signals to a digital data signal,

means for processing said digital data signal to generate a power spectral estimate thereof by means of maximum entropy estimation,

means to calculate a phase estimate of said digital data signal using said power spectral estimate, and calculating an estimate of the delay time of said echo signals being the travel time within said medium to said interface boundary,

means to calculate the distance to said interfaced boundary using said estimate of the delay time so as to locate the echo source of said echo waveforms,

means for processing said digital data signal to account for the sources of said echo waveforms wherein the contribution of predetermined echo waveforms is minimized.

* * * * *

US006104761A

United States Patent [19]

McCallister et al.

[11] **Patent Number: 6,104,761**

[45] **Date of Patent: Aug. 15, 2000**

[54] **CONSTRAINED-ENVELOPE DIGITAL-COMMUNICATIONS TRANSMISSION SYSTEM AND METHOD THEREFOR**

[75] Inventors: **Ronald D. McCallister**, Scottsdale; **Bruce A. Cochran**, Mesa; **Bradley P. Badke**, Chandler, all of Ariz.

[73] Assignee: **SiCom, Inc.**, Scottsdale, Ariz.

[21] Appl. No.: **09/143,230**

[22] Filed: **Aug. 28, 1998**

[51] **Int. Cl.**[7] **H04K 1/02**; H04L 25/03; H04L 25/49

[52] **U.S. Cl.** **375/296**; 375/261; 375/285; 375/298; 332/103

[58] **Field of Search** 375/295, 296, 375/298, 300, 302, 308, 377, 285, 259, 261, 268, 271, 279, 281, 284, 286, 291; 332/103, 144, 149

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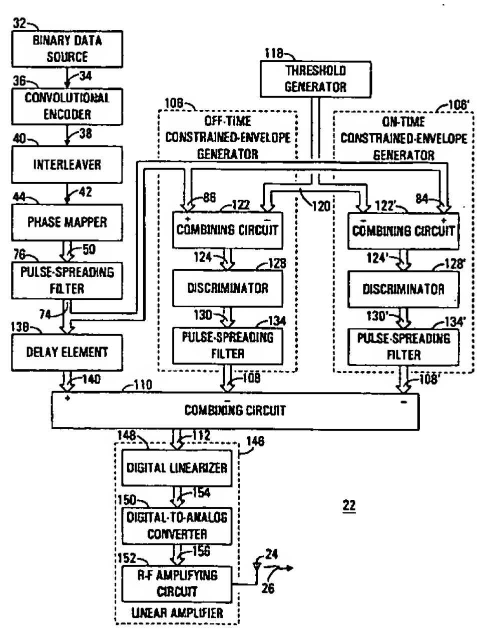
Primary Examiner—Chi H. Pham
Assistant Examiner—Jean B. Corrielus
Attorney, Agent, or Firm—Lowell W. Gresham, Esq.; Jordan M. Meschkow, Esq.; Meschkow & Gresham, P.L.C.

[57] **ABSTRACT**

A constrained-envelope digital-communications transmitter circuit **(22)** in which a binary data source **(32)** provides an input signal stream **(34)**, a phase mapper **(44)** maps the input signal stream **(34)** into a quadrature phase-point signal stream **(50)** having a predetermined number of symbols per unit baud interval **(64)** and defining a phase point **(54)** in a phase-point constellation **(46)**, a pulse-spreading filter **(76)** filters the phase-point signal stream **(50)** into a filtered signal stream **(74)**, a constrained-envelope generator **(106)** generates a constrained-bandwidth error signal stream **(108)** from the filtered signal stream **(74)**, a delay element **(138)** delays the filtered signal stream **(74)** into a delayed signal stream **(140)** synchronized with the constrained-bandwidth error signal stream **(108)**, a complex summing circuit **(110)** sums the delayed signal stream **(140)** and the constrained-bandwidth error signal stream **(108)** into a constrained-envelope signal stream **(112)**, and a substantially linear amplifier **(146)** amplifies the constrained-envelope signal stream **(112)** and transmits it as a radio-frequency broadcast signal **(26)**.

29 Claims, 4 Drawing Sheets



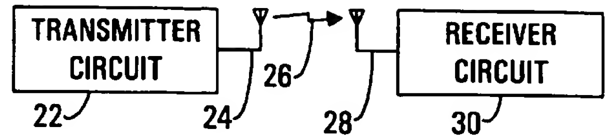
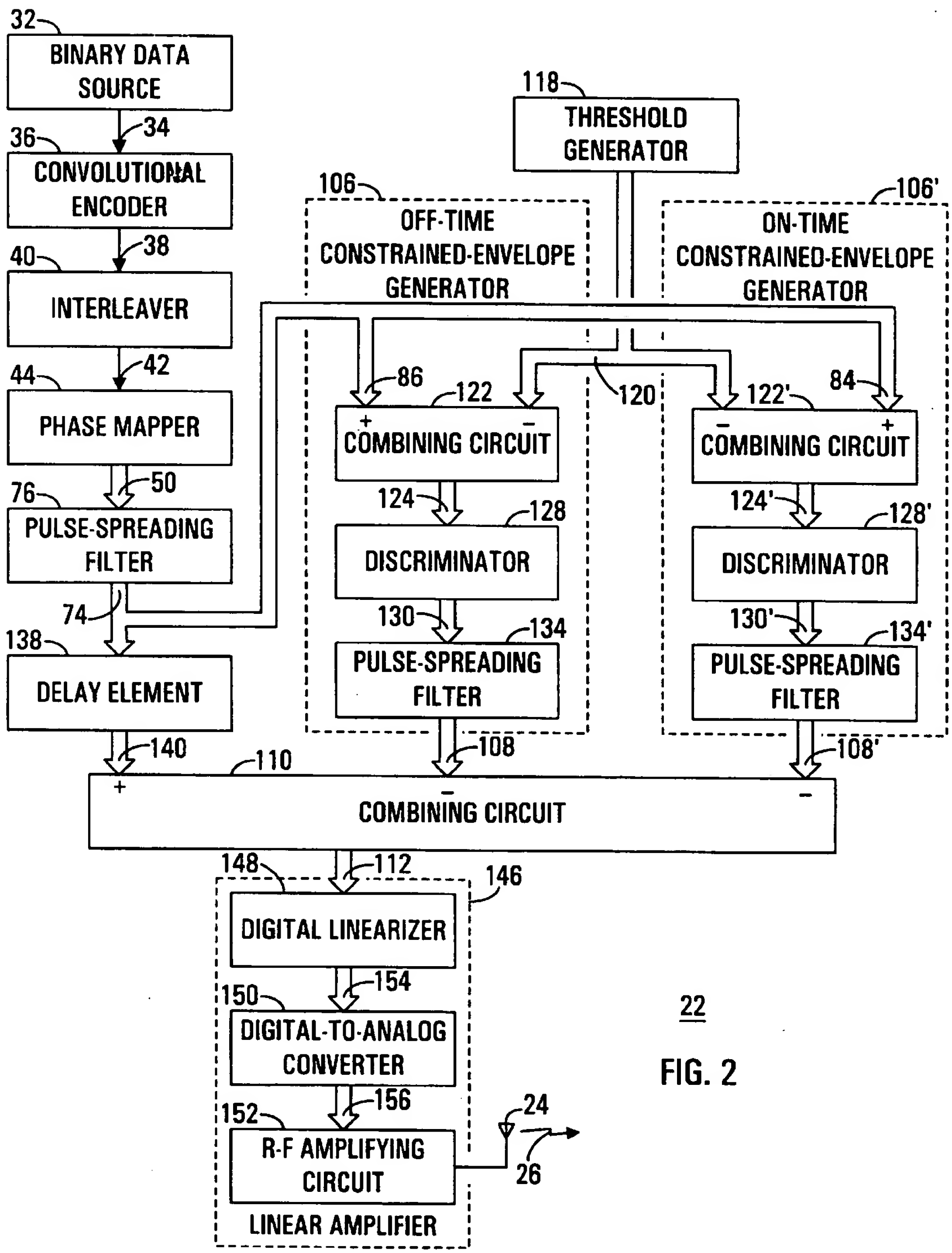

TRANSMITTER CIRCUIT
RECEIVER CIRCUIT
22
24
26
28
30
20

FIG. 1


BINARY DATA SOURCE
CONVOLUTIONAL ENCODER
INTERLEAVER
PHASE MAPPER
PULSE-SPREADING FILTER
DELAY ELEMENT
THRESHOLD GENERATOR
OFF-TIME CONSTRAINED-ENVELOPE GENERATOR
ON-TIME CONSTRAINED-ENVELOPE GENERATOR
COMBINING CIRCUIT
DISCRIMINATOR
PULSE-SPREADING FILTER
COMBINING CIRCUIT
DISCRIMINATOR
PULSE-SPREADING FILTER
COMBINING CIRCUIT
DIGITAL LINEARIZER
DIGITAL-TO-ANALOG CONVERTER
R-F AMPLIFYING CIRCUIT
LINEAR AMPLIFIER
32
34
36
38
40
42
44
50
76
74
138
140
110
118
106
106'
86
122
120
122'
84
124
128
124'
128'
130
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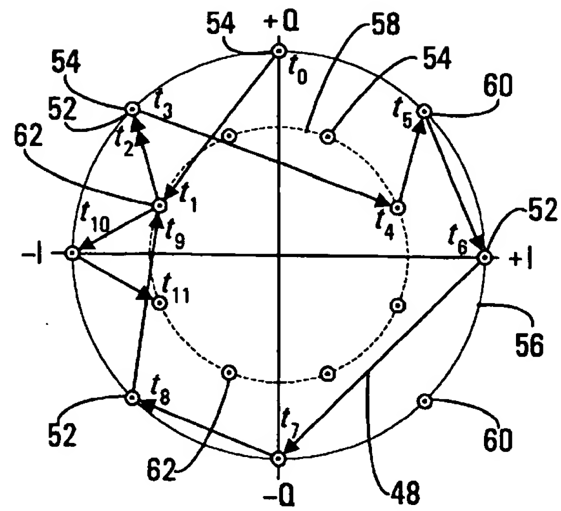
130'
134'
108
108'
148
112
146
150
154
152
156
24
26
22

FIG. 2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-Q
-I
+I
54
58
54
60
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52
62
52
56
52
62
48
60
t_0
t_1
t_2
t_3
t_4
t_5
t_6
t_7
t_8
t_9
t_{10}
t_{11}

46

FIG. 3

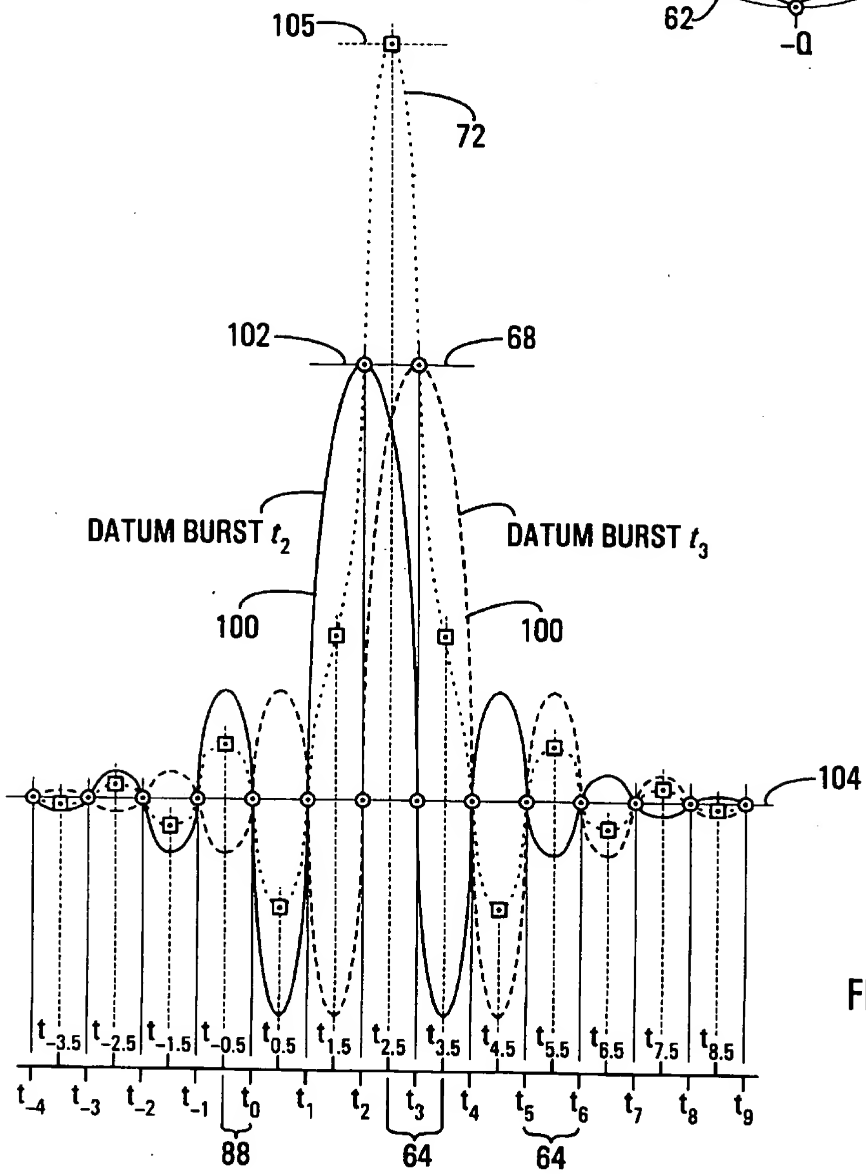
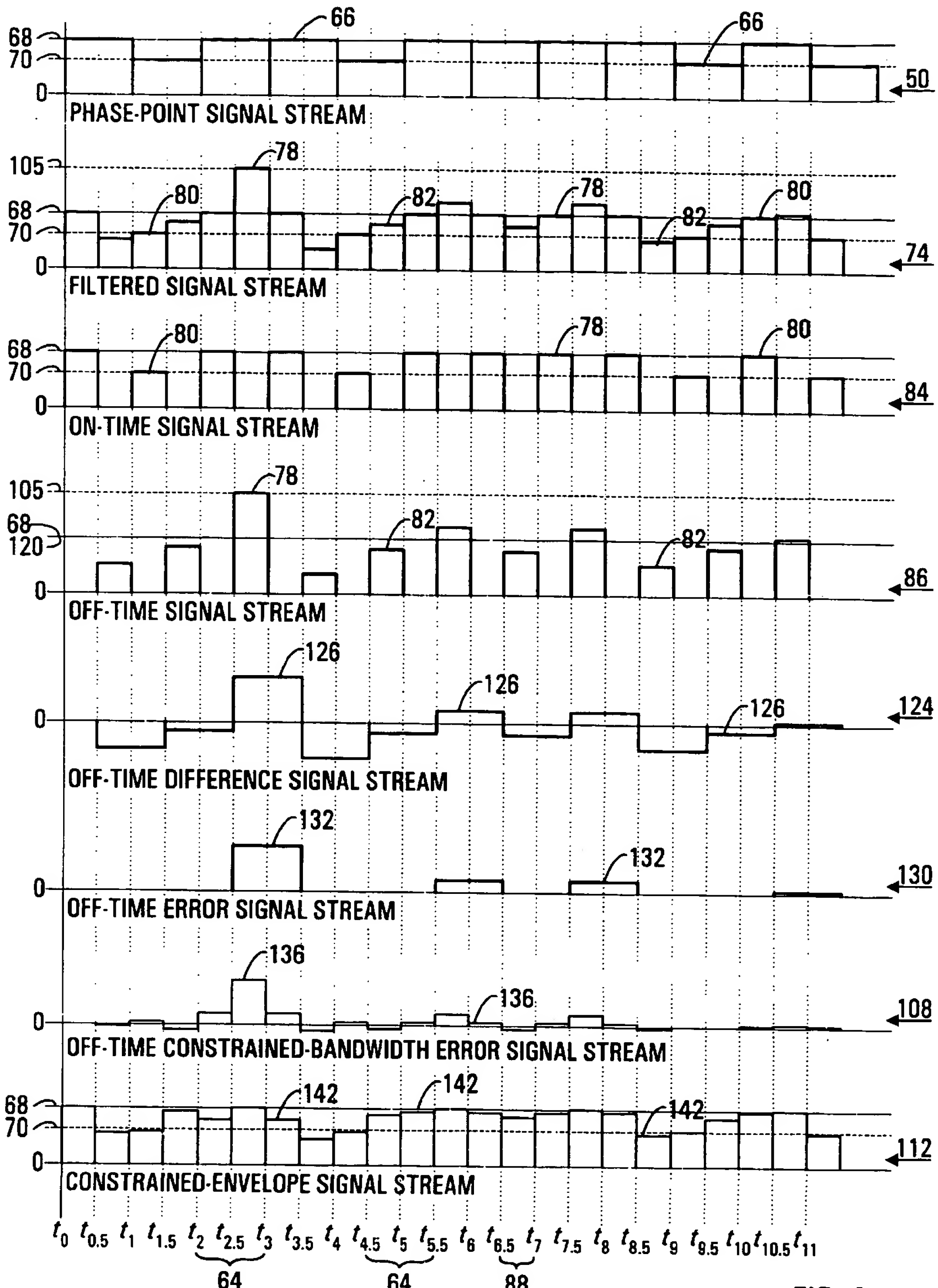
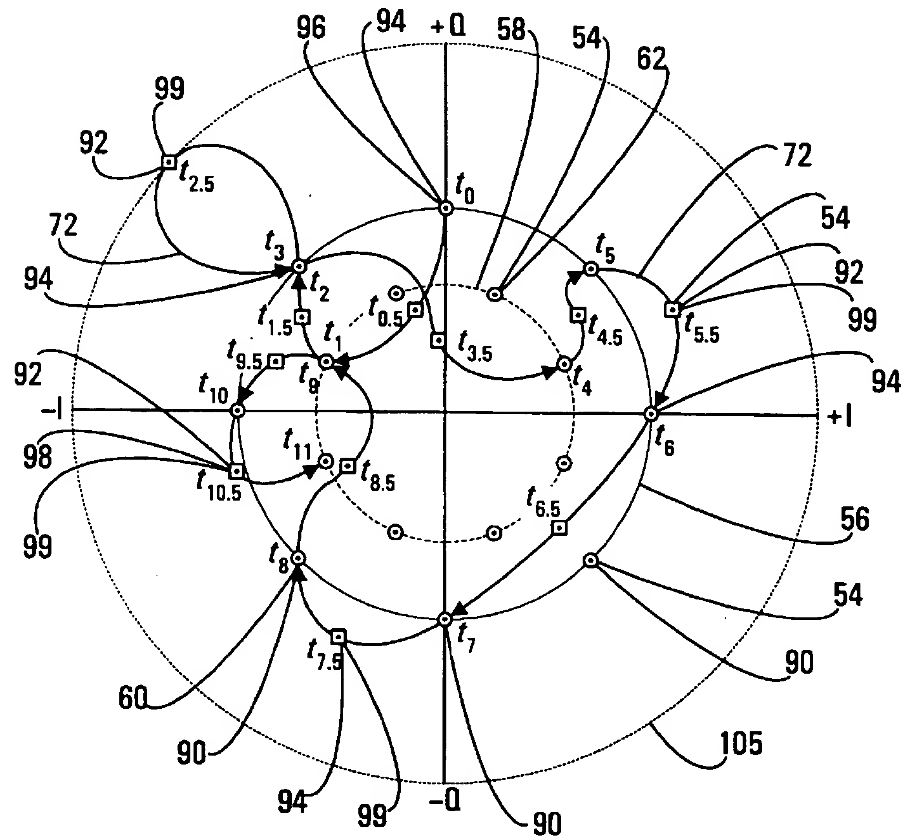

105
72
102
68
DATUM BURST t_2
DATUM BURST t_3
100
100
104
$t_{-3.5}$ $t_{-2.5}$ $t_{-1.5}$ $t_{-0.5}$ $t_{0.5}$ $t_{1.5}$ $t_{2.5}$ $t_{3.5}$ $t_{4.5}$ $t_{5.5}$ $t_{6.5}$ $t_{7.5}$ $t_{8.5}$
t_{-4} t_{-3} t_{-2} t_{-1} t_0 t_1 t_2 t_3 t_4 t_5 t_6 t_7 t_8 t_9
88
64
64

FIG. 6



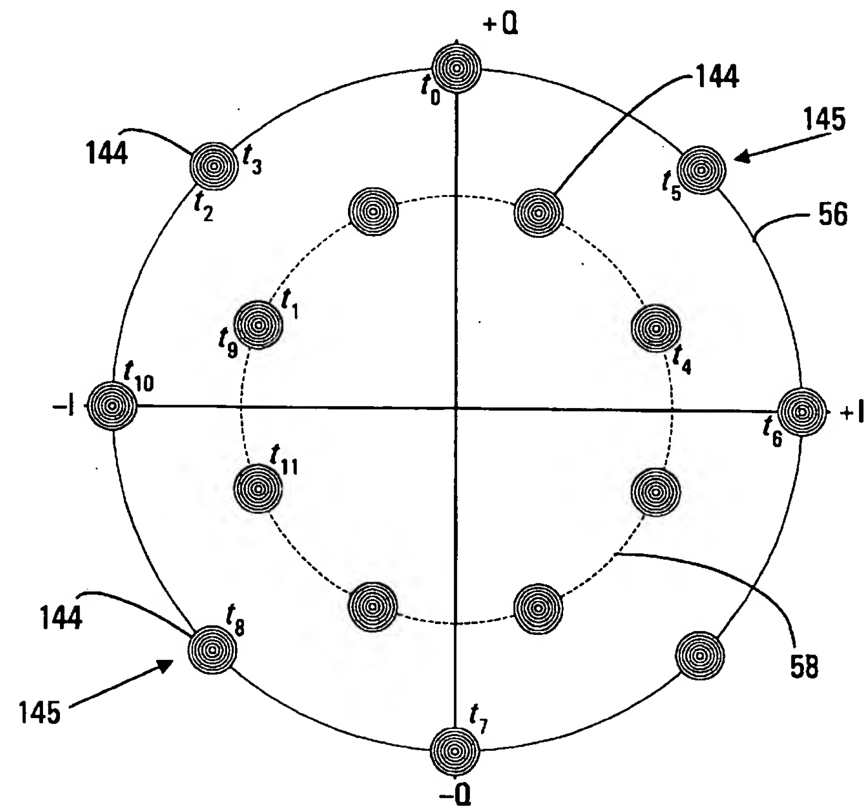
68
70
0
66
50
PHASE-POINT SIGNAL STREAM
105
78
80
82
74
FILTERED SIGNAL STREAM
84
ON-TIME SIGNAL STREAM
120
86
OFF-TIME SIGNAL STREAM
126
124
OFF-TIME DIFFERENCE SIGNAL STREAM
132
130
OFF-TIME ERROR SIGNAL STREAM
136
108
OFF-TIME CONSTRAINED-BANDWIDTH ERROR SIGNAL STREAM
142
112
CONSTRAINED-ENVELOPE SIGNAL STREAM
t_0 $t_{0.5}$ t_1 $t_{1.5}$ t_2 $t_{2.5}$ t_3 $t_{3.5}$ t_4 $t_{4.5}$ t_5 $t_{5.5}$ t_6 $t_{6.5}$ t_7 $t_{7.5}$ t_8 $t_{8.5}$ t_9 $t_{9.5}$ t_{10} $t_{10.5}$ t_{11}
64
64
88

FIG. 4



46'

FIG. 5



46"

FIG. 7

CONSTRAINED-ENVELOPE DIGITAL-COMMUNICATIONS TRANSMISSION SYSTEM AND METHOD THEREFOR

TECHNICAL FIELD OF THE INVENTION

The present invention relates generally to the field of digital communications. More specifically, the present invention relates to the field of constrained-envelope digital transmitter circuits.

BACKGROUND OF THE INVENTION

A wireless digital communications system should ideally refrain from using any portion of the frequency spectrum beyond that actually required for communications. Such a maximally efficient use of the frequency spectrum would allow the greatest number of communications channels per given spectrum. In the real-world, however, some spectral regrowth (i.e., increase in spectral bandwidth) is inevitable due to imperfect signal amplification.

In wireless communication systems various methodologies have been used to minimize spectral regrowth. Some conventional methodologies utilize complex digital signal processing algorithms to alter a digitally modulated transmission signal in some manner conducive to minimal spectral regrowth. Such complex algorithmic methodologies are well suited to low-throughput applications, i.e., those less than 0.5 Mbps (megabits per second), such as transmission of vocoder or other audio data. This is because the low throughput rate allows sufficient time between symbols for the processor to perform extensive and often repetitive calculations to effect the required signal modification. Unfortunately, high-throughput applications, i.e., those greater than 0.5 Mbps, such as the transmission of high-speed video data, cannot use complex processing algorithms because the processing power required to process the higher data rate is impractical.

A digital signal processing methodology may be used with the transmission of burst signals. With burst transmissions, the interstitial time between bursts may be used to perform the necessary complex computations based upon an entire burst. This methodology is not practical when continuous (as opposed to burst) transmission is used.

A conventional form of post-modulation pulse shaping to minimize spectral bandwidth utilizes some form of Nyquist-type filtration, such as Nyquist, root-Nyquist, raised cosine-rolloff etc. Nyquist-type filters are desirable as they provide a nearly ideal spectrally constrained waveform and negligible inter-symbol interference. This is achieved by spreading the datum for a single constellation phase point over many unit baud intervals in such a manner that the energy from any given phase-point datum does not interfere with the energy from preceding and following phase-point data at the appropriate baud-interval sampling instants.

The use of Nyquist-type filtration in a transmission circuit produces a filtered signal stream containing a pulse waveform with a spectrally constrained waveform. The degree to which a Nyquist-type pulse waveform is constrained in bandwidth is a function of the excess bandwidth factor, α. The smaller the value of α, the more the pulse waveform is constrained in spectral regrowth. It is therefore desirable to have the value of α as small as possible. However, as the value of α is decreased, the ratio of the spectrally constrained waveform magnitude to the spectrally unconstrained waveform magnitude is increased. The spectrally unconstrained waveform is the waveform that would result if no action were taken to reduce spectral regrowth. Typical designs use a values of 0.15 to 0.5. For an exemplary a value of 0.2, the magnitude of the spectrally constrained waveform is approximately 1.8 times that of the unconstrained waveform. This means that, for a normalized spectrally unconstrained waveform magnitude power of 1.0, the transmitter output amplifier must actually be able to provide an output power of 3.24 (1.8^2) to faithfully transmit the spectrally constrained waveform. This poses several problems.

When the transmitter output amplifier is biased so that the maximum spectrally unconstrained waveform (1.0 normalized) is at or near the top of the amplifier's linear region, all "overpower" will be clipped as the amplifier saturates. Such clipping causes a marked increase in spectral regrowth, obviating the use of Nyquist-type filtration.

When the transmitter output amplifier is biased so that the maximum spectrally constrained waveform (1.8 normalized) is at or near the top of the amplifier's linear region, the spectrally unconstrained waveform is at only 56 percent (i.e., $1/_{1.8}$) of the amplifiers peak linear power. This results in an inefficient use of the output amplifier.

Also, the biasing of the transmitter output amplifier so that the spectrally constrained waveform is at or near the top of the amplifier's linear region requires that the output amplifier be of significantly higher power than that required for the transmission of a spectrally unconstrained waveform. Such a higher-power amplifier is inherently more costly than its lower-power counterparts.

SUMMARY OF THE INVENTION

It is an advantage of the present invention that a circuitry and a methodology are provided that allow a transmitter output amplifier to be biased so that the spectrally unconstrained waveform is at or near the top of the amplifier's linear region without incurring clipping of a spectrally constrained waveform.

It is another advantage of the present invention that a circuitry and methodology are provided that allow a spectrally constrained waveform to have approximately the same magnitude as the spectrally unconstrained waveform without effecting a significant increase in spectral regrowth.

It is another advantage of the present invention that a circuitry and methodology are provided which allow a spectrally constrained waveform to be utilized with a continuous transmission scheme.

It is another advantage of the present invention that a circuitry and methodology are provided which allow efficient use of a transmitter output amplifier, thus allowing higher power output for a given output amplifier and a given bandwidth constraint than was previously feasible.

It is another advantage of the present invention that a circuitry and methodology are provided which allow efficient use of a transmitter output amplifier, which allows allowing a lower-power amplifier to be used for achieving given bandwidth constraints than was previously feasible, thus effecting a significant saving in the cost thereof.

These and other advantages are realized in one form by a constrained-envelope digital communications transmitter circuit. The transmitter circuit has a pulse-spreading filter configured to receive a quadrature phase-point signal stream of digitized quadrature phase points and produce a filtered signal stream, which filtered signal stream exhibits energy corresponding to each phase point spread throughout a plurality of baud intervals. The transmitter circuit also has a constrained-envelope generator coupled to the pulse-spreading filter and configured to produce a constrained-

bandwidth error signal stream. The transmitter circuit also has a combining circuit coupled to the pulse-spreading filter and to the constrained-envelope generator, which combining circuit is configured to combine the filtered signal stream and the constrained-bandwidth error signal stream to produce a constrained-envelope signal stream. The transmitter circuit also has a substantially linear amplifier with an input coupled to the combining circuit.

These and other advantages are realized in another form by a method for the transmission of a constrained-envelope communications signal in a digital communications system. The transmission method includes the step of filtering a quadrature phase-point signal stream to produce a filtered signal stream, which filtering step spreads energy from each phase point over a plurality of baud intervals. The transmission method also includes the step of generating a constrained-bandwidth error signal stream from the filtered signal stream and a threshold signal. The transmission method also includes the step of combining the filtered signal stream and the constrained-bandwidth error signal stream to produce a constrained-envelope signal stream. The transmission method also includes the step of linearly amplifying the constrained-envelope signal stream to produce the constrained-envelope communications signal. The transmission method also includes the step of transmitting the constrained-envelope communications signal.

BRIEF DESCRIPTION OF THE DRAWINGS

A more complete understanding of the present invention may be derived by referring to the detailed description and claims when considered in connection with the Figures, wherein like reference numbers refer to similar items throughout the Figures, and:

FIG. 1 depicts a simplified block diagram of a digital communications system in accordance with a preferred embodiment of the present invention;

FIG. 2 depicts a block diagram of a constrained-envelope digital communications transmitter circuit in accordance with a preferred embodiment of the present invention;

FIG. 3 depicts a 16-P-APSK constellation illustrating a locus of a quadrature phase-point signal stream over twelve exemplary consecutively mapped phase points in accordance with a preferred embodiment of the present invention;

FIG. 4 depicts a plurality of signal streams in accordance with a preferred embodiment of the present invention;

FIG. 5 depicts the phase-point constellation of FIG. 3 illustrating an exemplary locus of a filtered signal stream over the twelve consecutively mapped phase points of FIG. 3 in accordance with a preferred embodiment of the present invention;

FIG. 6 depicts a pair of Nyquist-type data bursts in accordance with a preferred embodiment of the present invention; and

FIG. 7 depicts a noise-influenced constellation illustrating constrained-envelope phase-point probabilities of the phase points of the constellation of FIG. 3 in accordance with a preferred embodiment of the present invention.

DETAILED DESCRIPTION OF THE PREFERRED EMBODIMENTS

FIG. 1 depicts a simplified block diagram of a digital communications system 20 and FIG. 2 depicts a block diagram of a constrained-envelope digital communications transmitter circuit 22 in accordance with a preferred embodiment of the present invention. The following discussion refers to FIGS. 1 and 2.

Digital communications system 20, as depicted in FIG. 1, includes a transmitter circuit 22 and a transmitter antenna 24 together configured to modulate and transmit a radio-frequency (RF) broadcast signal 26 to a receiver antenna 28 and a receiver circuit 30, together configured to receive and demodulate RF broadcast signal 26. Those skilled in the art will appreciate that the embodiment of system 20 depicted is a simplistic one for purposes of discussion only. In normal use, system 20 would likely be a complex system consisting of many more components and broadcast signals. It will be appreciated that the use of such a complex communications system for system 20 in no way departs from the spirit of the present invention or the scope of the appended claims.

Transmitter circuit 22 has a binary data source 32 providing a binary input signal stream 34. Binary data source 32 may be any circuitry, device, or combination thereof producing input signal stream 34. Input signal stream 34 is made up of binary data that may be pre-encoded in any desired manner. That is, input signal stream 34 may be made up of data that has no encoding, concatenated encoding, Reed-Solomon block encoding, or any other form of encoding desired for or required of the communications scheme in use.

In the preferred embodiment, input signal stream 34 is a steam of continuous data (as contrasted with burst data) passing to an input of a convolutional encoder 36. Convolutional encoder 36 convolutionally encodes (e.g., Viterbi encodes) input signal stream 34 into an encoded signal stream 38. The use of convolutional encoder 36 in transmitter circuit 22 and a like convolutional decoder (not shown) in receiver circuit 30 significantly reduces the error rate of the overall signal in a manner well understood by those skilled in the art. However, convolutional encoder 36 may be omitted.

Interleaver 40 temporally decorrelates encoded signal stream 38 to produce an interleaved signal stream 42. That is, the symbols making up the binary signal stream are temporally decorrelated (i.e., separated) in transmitter circuit 22 and temporally correlated in receiver circuit 30. This is done so that correlated errors produced by downstream transmitter components, discussed hereinbelow, will then be decorrelated through a complimentary de-interleaver located in receiver circuit 30 before convolutional decoding in receiver circuit 30.

In the preferred embodiment, interleaved signal stream 42 passes to an input of a phase mapper 44. Those skilled in the art will appreciate that interleaver 40 is not desired in all embodiments of transmitter circuit 22, for example when convolutional encoder 36 is omitted. When interleaver 40 is omitted, encoded signal stream 38 is passed directly to the input of phase mapper 44. When both convolutional encoder 36 and interleaver 40 are omitted, binary input signal stream passes directly to the input of phase mapper 44.

FIG. 3 depicts a sixteen phase-point polar amplitude and phase shift keying (16-P-APSK) constellation 46 illustrating a locus 48 of a quadrature phase-point signal stream 50 (FIG. 2) over twelve exemplary sequential phase points 52 in accordance with a preferred embodiment of the present invention. The following discussion refers to FIGS. 2 through 3.

Phase mapper 44 maps symbols (i.e., binary data units) present in interleaved signal stream 42, encoded signal stream 38, or input signal stream 34, into phase points 54 in phase-point constellation 46. While constellation 46 is depicted in FIG. 3 as a 16-P-APSK constellation, those skilled in the art will appreciate that the circuitry and

methodology of the present invention may be applied to all forms of constellations. The present invention is especially beneficial when used with constellations having rings of different magnitudes, i.e., amplitude and phase-shift keying (APSK) constellations. This is true because APSK constellations, requiring amplitude modulation of the signal, desirably use linear amplifiers to reproduce that amplitude modulation.

Each phase point **54** in constellation **46** represents a plurality, in this example four, of symbols. The values of the symbols in a given phase point **54** determine the location of that phase point **54** within constellation **46** in a manner well known to those skilled in the art.

Each quadrature phase point **54** may be thought of as having a vector value expressed as I,Q in the Cartesian coordinate system, where I is the in-phase (abscissa) value and Q is the quadrature (ordinate) value of the vector, or expressed as M,ϕ in the polar coordinate system, where M is the magnitude and ϕ is the phase angle of the vector. In this discussion, the M,ϕ designation will be used throughout, as the vector magnitude is the most discussed vector component.

In the exemplary 16-P-APSK constellation **46** of FIG. **3**, each phase point **54** resides upon an outer ring **56** or an inner ring **58**. Phase-points **54** residing upon outer ring **56** are outer-ring or maximum-magnitude phase points **60**. That is, outer-ring phase points **60** have a maximum magnitude (maximum value of M) as represented by the radius of outer ring **56**. For purposes of discussion, the magnitudes of outer-ring phase points **60** are normalized to 1.00.

Inner-ring phase points **62**, i.e., those phase points **54** residing upon inner ring **58**, have a lesser magnitude as represented by the radius of inner ring **58**. For the exemplary 16-P-APSK constellation **46** depicted in FIG. **3**, the magnitudes of inner-ring phase points **62** may desirably be approximately 0.63 when outer-ring phase point **60** magnitudes are normalized to 1.00.

FIG. **4** depicts a plurality of signal streams, in accordance with a preferred embodiment of the present invention. The following discussion refers to FIGS. **2** through **4**.

The output of phase mapper **44** is phase-point signal stream **50**. Phase mapper **44** processes one phase point **54** per unit baud interval **64**. That is, phase-point signal stream **50** consists of a series of consecutive phase-point pulses **66**, each of which represents one phase point **54**, whose leading edges are one unit baud interval **64** apart. Those skilled in the art will appreciate that other embodiments of phase-point signal stream **50** are equally valid, that the embodiment utilized is dependent upon the circuitry producing and processing phase-point signal stream **50**, and that the use of other embodiments of this or any other signal stream does not depart from the spirit of the present invent nor the scope of the appended claims.

FIGS. **3** and **4** illustrate a series of twelve exemplary sequential phase points **52**, representative of a random data stream processed by transmitter circuit **22** (FIG.**2**). These twelve exemplary phase points **52** reside at temporally consecutive locations labeled t_0, t_1, t_2, t_3, t_4, t_5, t_6, t_7, t_8, t_9, t_{10}, and t_{11}. These labels represent sequential integral times at unit baud intervals **64**, i.e., integral-baud times, and indicate the leading-edge times of phase-point pulses **66**. For purposes of simplification within this discussion, any occurrence at time t_N shall be referred to as "occurrence tN". For example, an exemplary phase point **52** occurring at time t_2 shall be referred to as phase point t_2, and the associated phase-point pulse **66** whose leading edge occurs at time t_2 shall be referred to as phase-point-signal pulse t_2. In other words, at time t_2, phase point t_2 is clocked and phase-point-signal pulse t_2 begins. One unit baud interval **64** later, at time t_3, phase point t_3 is clocked and phase-point pulse t_3 begins. This process continues indefinitely, with twelve exemplary phase points t_0 through t_{11} depicted in FIG. **3** and twelve corresponding phase-point-signal pulses t_0 through t_{11} depicted in phase-point signal stream **50** of FIG. **4**.

Table 1 below illustrates the magnitudes for phase-point-signal pulses to through t_{11}.

TABLE 1

Phase-Point Pulse Magnitudes

Phase-Point-Signal Pulse	Magnitude
t_0	Outer-Ring 68
t_1	Inner-Ring 70
t_2	Outer-Ring 68
t_3	Outer-Ring 68
t_4	Inner-Ring 70
t_5	Outer-Ring 68
t_6	Outer-Ring 68
t_7	Outer-Ring 68
t_8	Outer-Ring 68
t_9	Inner-Ring 70
t_{10}	Outer-Ring 68
t_{11}	Inner-Ring 70

Phase point t_0 is an outer-ring phase point **60**. Phase-point-signal pulse to therefore has an outer-ring magnitude **68**. In like manner, phase point t_1 is an inner-ring phase point **62** and phase-point-signal pulse t_1 has an inner-ring magnitude **70**.

Phase-point signal stream **50** effects locus **48** through constellation **46**. Locus **48** coincides with the location of each exemplary phase point t_0 through t_{11} in turn at unit baud intervals **64**. In FIG. **3**, locus **48** is depicted as effecting a minimum distance (straight line) path between adjacent exemplary phase points **52**. Those skilled in the art will appreciate that locus **48** is so depicted solely for the sake of simplicity, and that in actual practice, locus **48** instantly jumps or snaps between exemplary phase points **52** in a discontinuous manner.

FIG. **5** depicts an expanded phase-point constellation **46'** illustrating a locus **72** of a filtered signal stream **74** (FIG. **2**) over twelve exemplary sequential phase points **52** in accordance with a preferred embodiment of the present invention. The following discussion refers to FIGS. **2** through **5**.

In the preferred embodiment, phase-point signal stream **50** passes to the input of a pulse-spreading filter **76**, preferably realized as a Nyquist-type filter, such as a Nyquist, root-Nyquist, raised cosine-rolloff, etc., filter. Pulse-spreading filter **76** filters phase-point signal stream **50** into filtered signal stream **74**, depicted in FIG. **5**. In orthogonal frequency division multiplex (OFDM) systems, also known as multitone modulation (MTM) systems, pulse-spreading filter **76** may be implemented using a transmultiplexer or equivalent circuitry.

In accordance with Shannon's theory, well known to those skilled in the art, pulse-spreading filter **76** produces at least two (only two in the preferred embodiment) output filtered-signal pulses **78**, i.e., complex samples of filtered signal stream **74**, for each input phase-point pulse **66** received. This is demonstrated in FIG. **4** where filtered signal stream **74** possesses two filtered-signal pulses **78** per unit baud interval **64**. In the preferred embodiment, filtered-signal pulses **78**

consist of alternating on-time pulses **80**, i.e., samples of filtered signal stream at integral unit baud intervals **64**, and off-time pulses **82**, i.e., samples of filtered signal stream **74** between integral unit baud intervals. In effect, filtered signal stream **74** is made up of two interleaved data streams, an on-time signal stream **84** and an off-time signal stream **86**.

On-time signal stream **84** is substantially a version of phase-point signal stream **50**, wherein each phase-point pulse **66** has been reduced in duration from one unit baud interval **64** to a half-unit baud interval **88** to become on-time pulse **80** while maintaining substantially the same relative leading-edge time. That is, filtered-signal pulse to has substantially the same magnitude and substantially the same leading edge time as phase-point pulse to with approximately one-half the duration. Of course, those skilled in the art will appreciate that signal streams **74** and **84** may be delayed from signal stream **50** by a delay imposed by filter **76**.

The generation of both on-time pulses **80** and off-time pulses **82** by pulse-spreading filter **76** effectively populates expanded constellation **46'** (FIG. **5**) with on-time phase-points **90** (circles) and off-time phase points **92** (squares). The original phase points **54** of constellation **46** (FIG. **3**), i.e., the phase points carrying the intelligence to be communicated by transmitter circuit **22**, are on-time phase points **90** of expanded constellation **46'**.

Added to expanded constellation **46'** are off-time phase points **92**, with each off-time phase-point **92** occurring approximately midway in time between consecutive on-time phase points **90**. Therefore, exemplary sequential phase points **52** become exemplary filtered phase points **94**. Exemplary filtered phase points **94** are made up of alternating exemplary on-time filtered phase points **96** and exemplary off-time filtered phase points **98**, and reside at temporally consecutive locations labeled t_0, $t_{0.5}$, t_1, $t_{1.5}$, t_2, $t_{2.5}$, t3, $t_{3.5}$, t_4, $t_{4.5}$, t_5, $t_{5.5}$, t_6, $t_{6.5}$, t_7, $t_{7.5}$, t_8, $t_{8.5}$, t_9, $t_{9.5}$, t_{10}, $t_{10.5}$, and t_{11}, In FIG. **5**, exemplary on-time filtered phase points **96** are located at integral-baud times (t_0, t_1, t_2, etc.), whereas exemplary off-time filtered phase points **98** are located at fractional-baud (non-integral-baud) times ($t_{0.5}$,$t_{1.5}$, $t_{2.5}$, etc.).

The generation of off-time phase points **92** approximately midway in time between consecutive on-time phase points **90** causes filtered signal locus **72** to effect excursions having local peak magnitudes **99** greater than outer-ring magnitude **68**. Such excursions occur because the immediate position of locus **72** at any given instant in time is not only a result of those phase points **54** proximate that position, but of a plurality of phase points **54** both preceding and following that instant in time. That is, in the preferred embodiment, the determination of the position of locus **72** at time $t_{2.5}$ (i.e., coincident with off-time phase point $t_{2.5}$) is determined not only by the positions of phase points t_2 and t_3, but by the positions of numerous phase points **54** preceding phase point $t_{2.5}$ (i.e., phase points t_2, $t_{1.5}$, t_1, $t_{0.5}$, etc.) and the positions of numerous phase points **54** following phase point $t_{2.5}$ (i.e., phase points t_3, $t_{3.5}$, t_4, $t_{4.5}$, etc.)

This phenomenon is illustrated in FIG. **6**, which depicts a pair of Nyquist-type datum bursts **100** in accordance with a preferred embodiment of the present invention. The following discussion refers to FIGS. **2**, **4**, **5**, and **6**.

In the preferred embodiment, pulse-spreading filter **76** is realized as a Nyquist-type filter. Therefore, when a single phase-point pulse **66** is filtered by pulse-spreading filter **76**, that single pulse **66** is transformed into a Nyquist-type datum burst **100** extending over a plurality of unit baud intervals **64**. It is a property of Nyquist-type filters that datum burst **100** attains a datum-burst peak value **102** (i.e., a local peak magnitude) at the primary sampling time of the specific phase-point pulse **66** (i.e., at time t_2 for phase-point pulse t_2), and attains a zero datum-burst value **104** (i.e., is equal to zero) at integral unit baud intervals **64** preceding and following peak datum-burst value **102** (i.e., at times . . . , t_{-1}, t_0, t_1, and t_3, t_4, t_5, . . . , for phase point pulse t_2). In this manner, the energy of each pulse **78** is spread over a plurality of baud intervals **64** preceding and following the clocking instant (time t_2).

FIG. **6** illustrates Nyquist-type datum bursts **100** for phase-point pulses t_2 and t_3, with datum burst t_2 depicted as a solid line and datum burst t_3 depicted as a dashed line. As an example, it may be seen from FIG. **6** that at time t_2 the value of datum burst **12** is peak datum-burst value **102**. At every other time separated from time t_2 by an integral number of unit baud intervals **64**, the value of datum burst t_2 is zero. An analogous condition occurs for datum burst t_3.

The value of locus **72** is, at each moment in time, the sum of all datum bursts **100** at that moment. In the simplified two-datum-burst example of FIG. **6**, locus **72**, depicted by a dotted line, is the sum of datum burst t_2 and datum burst t_3. Since datum bursts t_2 and t_3 are zero at each integral time t_N except times t_2 and t_3, the value of locus **72** is also zero except at times t_2 and t_3, where it assumes the peak values of datum bursts t_2 and t_3, respectively.

The value of locus **72** at any instant in time between integral-baud times is the sum of the values of all datum bursts **100** at that instant. For example, in FIG. **6** where only two datum bursts **100** are considered, locus **72** has a value at time $t_{2.5}$ that is the sum of the values of datum bursts t_2 and t_3 at time $t_{2.5}$. Since datum bursts t_2 and t_3 both have significant positive values at time $t_{2.5}$, locus **72** has a value significantly greater than the maximum values of either datum burst t_2 or datum burst t_3.

Since locus **72** describes the sum of all datum bursts **100**, locus **72** is a function of the shape of the curves (FIG.**6**) describing those datum bursts **100**. That is, locus **72** is a function of a filtered-signal peak magnitude component of a filtered-signal complex digital value at any given point. The shape of the datum-burst curve is a function of the excess bandwidth factor, α, a design property of pulse-spreading filter **76**. The smaller the value of α, the more locus **72** may rise above the peak datum burst values **102** of adjacent datum bursts **100**. Typical designs of pulse-spreading filters **76** use α values of 0.15 to 0.5. For like-valued adjacent phase points **54** and an α value of 0.2, a maximum excursion magnitude **105** (i.e., the potential local peak magnitude **99** of locus **72**) is approximately 1.8 times the value of the maximum phase-point magnitude. That is, the magnitude of the constrained envelope is approximately 1.8 times that of the unconstrained envelope. In the preferred embodiment depicted in FIGS. **3**, **4**, and **6**, on-time phase points t_2 and t_3 are both outer-ring phase points **60** having a normalized outer-ring magnitude **68** of 1.00. Therefore, off-time phase point $t_{2.5}$ may have a normalized maximum excursion magnitude **105** of 1.8. This implies that transmitter circuit **22**, to faithfully transmit phase point $t_{2.5}$ without excessive distortion, and without the benefit of the present invention, would require an output power of 3.24 (1.8^2) times the power required to transmit phase point t_2 or t_3, which are representative of the highest magnitude intelligence-carrying phase points **54**. This represents an inefficient use of available power.

The following discussion refers to FIGS. **2**, **4**, and **5**.

Off-time signal stream **86**, a portion of filtered signal stream **74**, passes from an output of pulse-spreading filter **76**

to an input of an off-time constrained-envelope generator **106**. It is the task of off-time constrained-envelope generator **106** to produce an off-time constrained-bandwidth error signal stream **108** from off-time signal stream **86**. A complex summing or combining circuit **110** combines off-time constrained-bandwidth error signal stream **108** with a delayed version of filtered signal stream **74** (discussed below) to produce a constrained-envelope signal stream **112**. Constrained-envelope signal stream **112** is effectively filtered signal stream **74** with compensation for excursions of locus **72** with magnitudes greater than outer-ring magnitude **68**.

A quadrature threshold generator **118** generates a quadrature threshold signal **120**. In the preferred embodiment, threshold signal **120** is a steady-state, constant signal having a value approximately equal to outer-ring magnitude **68**. Threshold signal **120** is used to establish a reference with which off-time signal stream **86** is compared. Those skilled in the art will appreciate that threshold signal **120** may assume many forms and values in keeping with the methodology and circuitry incorporated in the comparison. The use of other forms and/or other values does not depart from the spirit of the present invention nor from the scope of the appended claims.

Threshold signal **120** and off-time signal stream **86** are combined in an off-time complex summing or combining circuit **122** to produce an off-time difference signal stream **124**. Off-time difference signal stream **124** is made up of a series of off-time difference pulses **126** whose values are the difference between the values of equivalent off-time pulses **82** and the value of threshold signal **120**. Since any given off-time pulse **82** may have a value greater than, equal to, or less than the value of threshold signal **120**, off-time difference signal stream **124** would normally be made up of a combination of off-time difference pulses **126** having positive, zero, and negative values.

Off-time difference signal stream **124** is passed to the input of an off-time discriminator **128** to produce an off-time error signal stream **130**. In the preferred embodiment, off-time error signal stream **130** is a variation of off-time difference signal stream **124** in which all off-time difference pulses **126** having positive values are passed unchanged as off-time error pulses **132** while all other off-time difference pulses **126** are passed as zero-value pulses (i.e., eliminated). In other words, off-time error signal stream **130** is formed from pulses, the timing of which coincide with excursions of locus **72** beyond outer-ring magnitude **68** and the magnitudes of which correspond to the degree to which locus **72** passes beyond outer-ring magnitude **68**.

Off-time error signal stream **130** is then passed to the input of an off-time pulse-spreading filter **134**. Off-time pulse-spreading filter **134** is substantially identical to first pulse-spreading filter **76**. That is, in the preferred embodiment, both pulse spreading filters **76** and **134** are realized as Nyquist-type filters with substantially identical transfer characteristics. Off-time pulse-spreading filter **134** produces off-time constrained-bandwidth error signal stream **108** and completes the action of off-time constrained-envelope generator **106**.

Within off-time constrained-envelope generator **106**, off-time pulse-spreading filter **134** receives one off-time error pulse **132** from off-time discriminator **128** per unit baud interval **64**. Off-time pulse-spreading filter **134** then transforms each off-time error pulse **132** into a Nyquist-type error burst (not shown) extending over a plurality of unit baud intervals. Since off-time pulse-spreading filter **134** is a Nyquist-type filter, each error burst attains an error-burst peak value (not shown) at the primary sampling time of the specific off-time error pulse **132** (i.e., at time $t_{2.5}$ for error pulse $t_{2.5}$), and attains a zero error-burst value (not shown) at integral unit baud intervals **64** preceding and following the peak error-burst value (i.e., at times . . . , $t_{-1.5}$, $t_{0.5}$, $t_{1.5}$, and $t_{3.5}$, $t_{4.5}$, $t_{5.5}$, . . . , for error pulse $t_{2.5}$). In this manner, the energy of each off-time constrained-envelope error pulse **136** is spread over a plurality of baud intervals **64** preceding and following the clocking instant (time $t_{2.5}$). This results in the conversion of off-time error signal stream **130** into off-time constrained-bandwidth error signal stream **108**. Off-time constrained-bandwidth error signal stream **108** is made up of off-time constrained-envelope error pulses **136**. This operation is essentially the same as the operation of pulse-spreading filter **76** in the conversion of phase-point signal stream **50** into filtered signal stream **74** described hereinabove.

Since off-time constrained-envelope error pulses **136** are derived from off-time pulses **82**, the error-burst peak and zero values occur approximately midway between integral baud times, i.e., at baud times $t_{0.5}$, $t_{1.5}$, $t_{2.5}$, etc., hence between datum-burst peak and zero values **102** and **104** of filtered signal stream **74**.

The production of off-time constrained-bandwidth error signal stream **108** completes the operation of off-time constrained envelope generator **106**.

Filtered signal stream **74** is also passed to the input of a delay element **138**. Delay element **138** produces delayed signal stream **140**, which is effectively filtered signal stream **74** delayed sufficiently to compensate for the propagation and other delays encountered in off-time constrained-envelope generator **106**, and particularly in off-time pulse-spreading filter **134**. In other words, delayed signal stream **140** is filtered signal stream **74** brought into synchronization with off-time constrained-bandwidth error signal stream **108**.

Combining circuit **110** combines filtered signal stream **74**, in the form of delayed signal stream **140**, and off-time constrained-bandwidth error signal stream **108** to reduce peak magnitude components of filtered signal stream **74**. A resultant constrained-envelope signal stream **112** is made up of a series of digital pulses **142** whose values are the difference between the values of corresponding filtered-signal pulses **78** and off-time constrained-envelope error pulses **136**. The result is a series of digital pulses **142** whose values do not appreciably exceed outer-ring magnitude **68** of expanded constellation **46'**.

In some embodiments of the present invention, certain of outer-ring phase points **60** may have magnitudes greater than outer-ring magnitude **68**, i.e., may be located beyond outer ring **56**. This condition may occur as a result of pulse-spreading filter **76** executing certain Nyquist-type functions well known to those skilled in the art. In such an embodiment, transmitter circuit **22** contains an on-time constrained envelope generator **106'** in addition to off-time constrained-envelope generator **106** discussed above.

On-time signal stream **84**, also a portion of filtered signal stream **74**, passes from an output of pulse-spreading filter **76** to an input of on-time constrained-envelope generator **106'**. It is the task of on-time constrained-envelope generator **106'** to produce an on-time constrained-bandwidth error signal stream **108'** from on-time signal stream **84**. Combining circuit **110** combines both off-time and on-time constrained-bandwidth error signal streams **108** and **108'** with the delayed version of filtered signal stream **74** (discussed below) to produce constrained-envelope signal stream **112**.

On-time constrained-envelope generator **106'** operates in a manner analogous with the operation of off-time constrained-envelope generator **106**. Threshold signal **120** and on-time signal stream **84** are combined in an on-time complex summing or combining circuit **122'** to produce an on-time difference signal stream **124'**. On-time difference signal stream **124'** is passed to the input of an on-time discriminator **128'** to produce an on-time error signal stream **130'**. On-time error signal stream **130'** is then passed to the input of an on-time pulse-spreading filter **134'**, which produces on-time constrained bandwidth error signal stream **108'**. Like off-time pulse-spreading filter **134**, on-time pulse-spreading filter **134'**, is substantially identical to first pulse-spreading filter **76**.

Since on-time constrained-envelope error pulses (not shown) are derived from on-time pulses **80**, the error-burst peak and zero values occur at integral baud times, i.e., at baud times t_1, t_2, t_3, etc., hence between datum-burst peak and zero values **102** and **104** of filtered signal stream **74**.

Combining circuit **110** combines filtered signal stream **74**, in the form of delayed signal stream **140**, with both off-time and on-time constrained-bandwidth error signal stream **108** and **108'** to reduce peak magnitude components of filtered signal stream **74**.

A side effect of this methodology is that locus **72** at integral unit baud intervals **64** adds a signal-dependent, baud-limited noise factor to the positions of phase points **54** in constellation **46** (FIG. **3**). This results in transmitter circuit **22** transmitting a "noise-influenced" phase-point constellation **46"**. In FIG. **7**, noise-influenced constellation **46"** is depicted illustrating constrained-envelope phase-point probabilities **144** of phase points **54** in accordance with a preferred embodiment of the present invention. The following discussion refers to FIGS. **2**, **3**, **5** and **7**.

Phase-point probabilities **144** reside in noise-influenced constellation **46"** exactly as phase points **54** reside in constellation **46**, i.e., in the same configuration with centers at the same locations. The actual location of a given transmitted phase point **145** within a given phase-point probability **144** is a function of a plurality of variable conditions and, although somewhat correlated, except in certain specialized cases, cannot readily be predicted. In effect, for a given phase point **54**, the resultant transmitted phase point **145** may be located anywhere within phase-point probability **144**, i.e., within an indeterminate area having a center coincident with the location of the original phase point **54**. The probability of transmitted phase point **145** being located at any specific position within that indeterminate area varies as an inverse function of the distance of that specific position from the location of the original phase point **54**.

For any given phase point **54**, the transmitted phase point **145** may be said to be proximate its idealized position within noise-influenced constellation **46"**. That is, a locus (not shown) of constrained-envelope signal stream **112** passes proximate the idealized positions of exemplary phase points t_0, t_1, t_2, etc., at the clocking instants in time.

The original phase points **54** of constellation **46**, as produced by phase mapper **44**, are on-time phase points **90** (circles) of expanded constellation **46'**. It is these on-time phase points **90** that carry the intelligence of RF broadcast signal **26** as ultimately transmitted. Off-time phase points **92** (squares) are by-products of pulse-spreading filter **76**, required to constrain spectral regrowth, and carry no intelligence. Phase-point probabilities **144** of noise-influenced constellation **46"** represent the resultant areas of probable locations of transmitted phase points **145** as derived from on-time phase points **90**. The centers of phase-point probabilities **144** occupy the same normalized locations within noise-influenced constellation **46"** as do on-time phase points **90** within expanded constellation **46'**.

The positional aberrations of transmitted phase points **145** relative to the corresponding on-time phase points **90** represent a degree of positional error. This positional error degrades the bit error rate and effects a detriment to transmission. The absence of off-time phase points **92** with a magnitude significantly greater than outer-ring magnitude **68** (FIG. **4**) in constrained-envelope signal stream **112**, however, allows an increase in power output for a given bandwidth and power amplifier that more than compensates for the position error of transmitted phase points **145**. A net improvement in performance results.

Referring back to FIG. **2**, the output of combining circuit **110**, constrained-envelope signal stream **112**, is passed to an input of a substantially linear amplifier **146**. Substantially linear amplifier **146** produces RF broadcast signal **26**, which is then broadcast via transmitter antenna **24**. In the preferred embodiment, substantially linear amplifier **146** is made up of a digital linearizer **148**, a digital-to-analog converter **150**, and a radio-frequency (RF) amplifying circuit **152**. Those skilled in the art will appreciate that substantially linear amplifier **146** may be realized in any of a plurality of different embodiments other than that described here, and that utilization of any of these different embodiment does not depart from the intent of the present invention nor the scope of the appended claims.

Within substantially linear amplifier **146**, digital linearizer **148** alters constrained-envelope signal stream **144** into a pre-distorted digital signal stream **154**. Pre-distorted digital signal stream **154** is made non-linear in just the right manner to compensate for non-linearities within digital-to-analog converter **150** and RF amplifying circuit **152**, hence linearizing substantially linear amplifier **146**.

Digital-to-analog converter **150** then converts pre-distorted digital signal stream **154** into an analog baseband signal **156**. Analog baseband signal **156** is then amplified by RF amplifying circuit **152** into RF broadcast signal **26** and transmitted via transmitter antenna **24**.

In summary, the present invention teaches a methodology and circuitry by which a transmitter circuit utilizing Nyquist-type filtration may produce a constrained envelope having a magnitude at or near the approximate unconstrained envelope magnitude of the desired constellation. This enables the transmitter output amplifier to be biased so that the maximum unconstrained envelope magnitude is at or near the top of the amplifier's linear region without incurring clipping of the constrained envelope transmissions. This in turn produces a more efficient output amplifier and effects an increase in the power output of a given output amplifier. Conversely, a lower power amplifier may be used to provide the same output power that was previously output. This effects a significant savings in output amplifier cost.

Although the preferred embodiments of the invention have been illustrated and described in detail, it will be readily apparent to those skilled in the art that various modifications may be made therein without departing from the spirit of the invention or from the scope of the appended claims.

What is claimed is:

1. A constrained-envelope digital communications transmitter circuit comprising:

a pulse-spreading filter configured to receive a quadrature phase-point signal stream of digitized quadrature phase

points and produce a filtered signal stream, said filtered signal stream exhibiting energy corresponding to each phase point spread throughout a plurality of unit baud intervals;

a constrained-envelope generator coupled to said pulse-spreading filter and configured to produce a constrained-bandwidth error signal stream;

a combining circuit coupled to said pulse-spreading filter and to said constrained-envelope generator, said combining circuit configured to combine said filtered signal stream and said constrained-bandwidth error signal stream to produce a constrained-envelope signal stream; and

a substantially linear amplifier having an input coupled to said combining circuit.

2. A digital communications transmitter circuit as claimed in claim 1 wherein said pulse-spreading filter is a Nyquist-type filter.

3. A digital communications transmitter circuit as claimed in claim 1 wherein said combining circuit is configured to combine said filtered signal stream and said constrained-bandwidth error signal stream to reduce a peak magnitude component of said filtered signal stream.

4. A digital communications transmitter circuit as claimed in claim 3 wherein said combining circuit is a complex summing circuit.

5. A digital communications transmitter circuit as claimed in claim 1 wherein:

said pulse-spreading filter is a first pulse-spreading filter;

said transmitter circuit additionally comprises a delay element coupled between said first pulse-spreading filter and said combining circuit; and

said constrained-envelope generator comprises a second pulse-spreading filter coupled to said combining circuit.

6. A digital communications transmitter circuit as claimed in claim 5 wherein:

said first pulse-spreading filter is configured so that each phase point is transformed into a Nyquist-type datum burst extending over a plurality of unit baud intervals, having a datum-burst peak value occurring in one of said plurality of unit baud intervals and datum-burst zero values occurring substantially at integral unit baud intervals away from said datum-burst peak value, so that said filtered signal stream in each unit baud interval substantially equals the sum of said Nyquist-type datum bursts from a plurality of phase points; and

said constrained-envelope generator is configured so that said second pulse-spreading filter receives error pulses, transforms each error pulse into a Nyquist-type error burst extending over a plurality of unit baud intervals, having an error-burst peak value occurring in one of said plurality of unit baud intervals and error-burst zero values occurring substantially at integral unit baud intervals away from said error-burst peak value, so that said constrained-bandwidth error signal stream in each unit baud interval substantially equals the sum of said Nyquist-type error bursts from a plurality of error pulses.

7. A digital communications transmitter circuit as claimed in claim 6 wherein said constrained-envelope generator is configured so that said Nyquist-type error bursts exhibit said error-burst peak values and said error-burst zero values at instances in time when said Nyquist-type datum bursts exhibit neither said datum-burst peak values nor said datum-burst zero values.

8. A digital communications transmitter circuit as claimed in claim 7 wherein said constrained-envelope generator is configured so that said error-burst peak values and said error-burst zero values occur approximately midway between said datum-burst peak values and said datum-burst zero values.

9. A digital communications transmitter circuit as claimed in claim 5 wherein said first and second pulse-spreading filters exhibit substantially equivalent transfer characteristics.

10. A digital communications transmitter circuit as claimed in claim 5 wherein:

said first pulse-spreading filter receives one quadrature phase point per unit baud interval and produces two complex samples of said filtered signal stream per unit baud interval;

said constrained-envelope generator evaluates one of said two complex samples of said filtered signal stream produced by said first pulse-spreading filter per unit baud interval; and

said second pulse-spreading filter receives one error pulse per unit baud interval and produces two complex samples of said constrained-envelope error-signal stream per unit baud interval.

11. A digital communications transmitter circuit as claimed in claim 1 wherein:

said filtered signal stream is a stream of complex digital values, with each of said complex digital values exhibiting a peak magnitude component; and

said constrained-envelope generator is configured to determine when ones of said peak magnitude components exceed a threshold value.

12. A digital communications transmitter circuit as claimed in claim 11 wherein:

said transmitter circuit additionally comprises a phase mapper coupled to said pulse-spreading filter and configured to select said digitized quadrature phase points from a phase-point constellation, said phase-point constellation having a maximum-magnitude phase point; and

said threshold value is a magnitude value approximately equal to a magnitude of said maximum-magnitude phase point.

13. A digital communications transmitter circuit as claimed in claim 1 additionally comprising an interleaver coupled to said phase mapper.

14. A digital communications transmitter circuit as claimed in claim 1 wherein:

said constrained-envelope generator is an off-time constrained-envelope generator;

said constrained-bandwidth error signal stream is an off-time constrained-bandwidth error signal stream;

said transmitter circuit additionally comprises an on-time constrained-envelope generator coupled to said pulse-spreading filter and configured to produce an on-time constrained-bandwidth error signal stream; and

said combining circuit is coupled to said pulse-spreading filter, to said off-time constrained-envelope generator, and to said on-time constrained-envelope generator, and said combining circuit is configured to combine said filtered signal stream, said off-time constrained-bandwidth error signal stream, and said on-time constrained-bandwidth error signal stream to produce said constrained-envelope signal stream.

15. A digital communications transmitter circuit as claimed in claim 1 wherein said substantially linear amplifier comprises;

a digital linearizer configured to pre-distort said constrained-envelope signal stream into a pre-distorted digital signal stream;

a digital-to-analog converter coupled to said digital linearizer and configured to produce an analog baseband signal from said pre-distorted digital signal stream; and

a radio-frequency amplifying circuit configured to generate a radio-frequency broadcast signal from said analog baseband signal.

16. In a digital communications system, a method for the transmission of a constrained-envelope communications signal, said transmission method comprising the steps of:

filtering a quadrature phase-point signal stream to produce a filtered signal stream, said filtering step spreading energy from each phase point in said filtered signal stream over a plurality of unit baud intervals;

generating a constrained-bandwidth error signal stream from said filtered signal stream and a threshold signal;

combining said filtered signal stream and said constrained-bandwidth error signal stream to produce a constrained-envelope signal stream;

linearly amplifying said constrained-envelope signal stream to produce said constrained-envelope communications signal; and

transmitting said constrained-envelope communications signal.

17. A transmission method as claimed in claim 16 wherein said combining step comprises the step of reducing a peak magnitude component of said filtered signal stream.

18. A transmission method as claimed in claim 16 wherein:

said generating step comprises the step of filtering an error signal stream having one error pulse per unit baud interval to produce said constrained-bandwidth error signal stream, said filtering step spreading energy from each error pulse in said error signal stream over a plurality of unit baud intervals;

said transmission method additionally comprises the step of delaying said filtered signal stream to produce a delayed signal stream; and

said combining step combines said delayed signal stream and said constrained-bandwidth error signal stream to produce said constrained-envelope signal stream.

19. A transmission method as claimed in claim 16 wherein:

said filtering step comprises the step of receiving one quadrature phase point per unit baud interval;

said filtering step additionally comprises the step of producing two complex samples of said filtered signal stream per unit baud interval;

said generating step comprises the step of evaluating one of said two complex samples of said filtered signal stream per unit baud interval to produce an error signal stream having one error pulse per unit baud interval; and

said generating step additionally comprises the step of filtering said error signal stream to produce said constrained-bandwidth error signal stream having two complex samples of said constrained-bandwidth error signal stream per unit baud interval.

20. A transmission method as claimed in claim 19 wherein said generating step additionally comprises the steps of:

providing said threshold signal; and

determining when ones of peak magnitude components of a stream of complex digital values of said filtered signal stream exceed a threshold value of said threshold signal.

21. A transmission method as claimed in claim 16 wherein:

said filtered signal stream includes two or more complex digital values per unit baud interval, said complex digital values in said filtered signal stream exhibiting local peak magnitudes; and

said generating step is configured so that said constrained-bandwidth error signal stream includes two or more complex values per unit baud interval, said complex values in said constrained-bandwidth error signal stream being responsive to said local peak magnitudes of said filtered signal stream so as to spread energy from selected ones of said local peak magnitudes over a plurality of unit baud intervals of said constrained-bandwidth error signal stream.

22. A transmission method as claimed in claim 16 wherein said transmitting step continuously transmits said constrained-envelope communications signal.

23. A constrained-envelope digital-communications transmitter circuit comprising:

a binary data source configured to provide a binary input signal stream;

a phase mapper coupled to said binary data source and configured to produce a quadrature phase-point signal stream, wherein said phase-point signal stream has a predetermined number of symbols per unit baud interval, said predetermined number of symbols defining a phase point in a phase-point constellation;

a Nyquist-type filter coupled to said phase mapper and configured to produce a filtered signal stream;

a constrained-envelope generator coupled to said Nyquist-type filter and configured to produce a constrained-bandwidth error signal stream;

a delay element coupled to said Nyquist-type filter and configured to produce a delayed signal stream synchronized with said constrained-bandwidth error signal stream;

a complex summing circuit coupled to said delay element and said constrained-envelope generator and configured to produce a constrained-envelope signal stream; and

a substantially linear amplifier coupled to said complex summing circuit and configured to produce a radio-frequency broadcast signal.

24. A digital-communications transmitter circuit as claimed in claim 23 wherein said Nyquist-type filter is a first Nyquist-type filter, said filtered signal stream includes a first filtered-signal data stream and a second filtered-signal data stream, and said complex summing circuit is a first complex summing circuit, wherein said transmitter circuit additionally comprises a quadrature threshold generator configured to-provide a threshold signal, said threshold signal having a threshold value, and wherein said constrained-envelope generator comprises:

a complex summing circuit coupled to said first Nyquist-type filter and said quadrature threshold generator and configured to produce a difference signal stream, wherein said difference signal stream is a stream of difference pulses having difference-pulse values of a first polarity and difference-pulse values of a second polarity;

a discriminator coupled to said complex summing circuit and configured to produce an error signal stream from said difference signal stream, wherein said error signal stream is a stream of error pulses substantially coinci-

dent with said difference pulses of said difference signal stream, and wherein, when ones of said difference pulses have said first-polarity difference-pulse values, said coincident error pulses have error-pulse values substantially equal to said first-polarity difference-pulse values, and when ones of said difference pulses have said second-polarity difference-pulse values, said coincident error pulses have error-pulse values substantially equal to zero; and

a second Nyquist-type filter coupled to said discriminator and configured to produce said constrained-bandwidth error signal stream.

25. A digital-communications transmitter circuit as claimed in claim **24** wherein said transmitter circuit additionally comprises:

a convolutional encoder coupled to said binary data source and configured to produce an encoded signal stream; and

an interleaver coupled to said convolutional encoder and configured to produce an interleaved signal stream by temporally decorrelating said encoded signal stream.

26. A digital-communications transmitter circuit as claimed in claim **24** wherein:

said filtered signal stream is a quadrature signal stream having a locus that passes proximate one of said phase points of said phase-point constellation at integral unit baud intervals;

said first filtered-signal data stream comprises on-time samples of said filtered signal stream, each of said on-time samples occurring substantially coincident ally with said passage of said filtered signal locus proximate one of said phase points of said phase-point constellation; and

said second filtered-signal data stream comprises off-time samples of said filtered signal stream wherein each of said off-time samples occurs between adjacent ones of said on-time samples.

27. A digital-communications transmitter circuit as claimed in claim **26** wherein each of said off-time samples occurs substantially midway between adjacent ones of said on-time samples.

28. A digital-communications transmitter circuit as claimed in claim **23** additionally comprising an interleaver coupled to said binary data source and configured to provide an interleaved signal stream.

29. A digital-communications transmitter circuit as claimed in claim **23** wherein said constellation is an amplitude and phase shift keying constellation.

* * * * *

UNITED STATES PATENT AND TRADEMARK OFFICE

CERTIFICATE OF CORRECTION

PATENT NO. : 6,104,761

DATED : 15 August 2000

INVENTOR(S) : McCallister et al

It is certified that error appears in the above-indentified patent and that said Letters Patent is hereby corrected as shown below:

In the Column 2, Line 1: please delete "a" and insert --α-- therefor.

In the Column 2, Line 1,: please delete "a" and insert --α-- therefor.

In the Column 5, Line 64,: please delete "tN" and insert --t_N--therefor.

In the Column 8, Line 15,: please delete "12" and insert --t_2--therefor.

In the Column 18, Line 5,: please delete "coincident ally" and insert --coincidentally--therefor.

Signed and Sealed this

Twenty-ninth Day of May, 2001

Attest:

Nicholas P. Godici

NICHOLAS P. GODICI

Attesting Officer

Acting Director of the United States Patent and Trademark Office